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Cyrusian

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(54) **TRUNCATION FOR THREE-LEVEL DIGITAL AMPLIFIER**

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(73) Assignee: **Marvell International Ltd.**, Hamilton (BM)

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(22) Filed: **Nov. 19, 2007**

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(51) **Int. Cl.**  
*H03M 5/08* (2006.01)

(52) **U.S. Cl.** ..... 341/53; 341/143; 341/152; 327/157; 330/10

(58) **Field of Classification Search** ..... 341/53, 341/152

See application file for complete search history.

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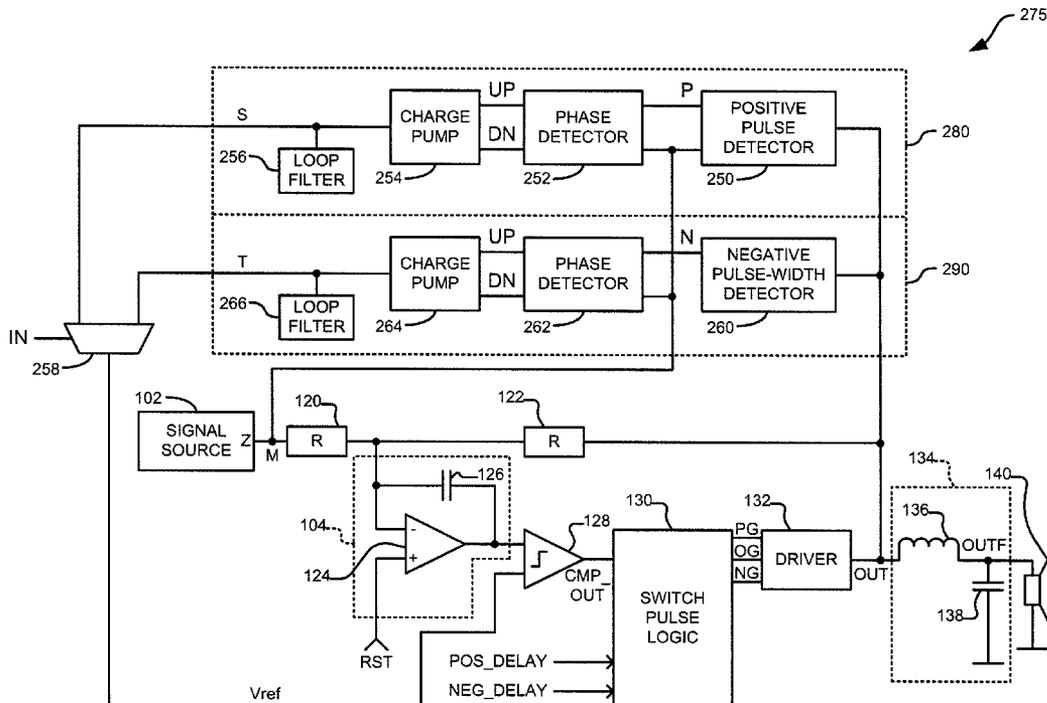
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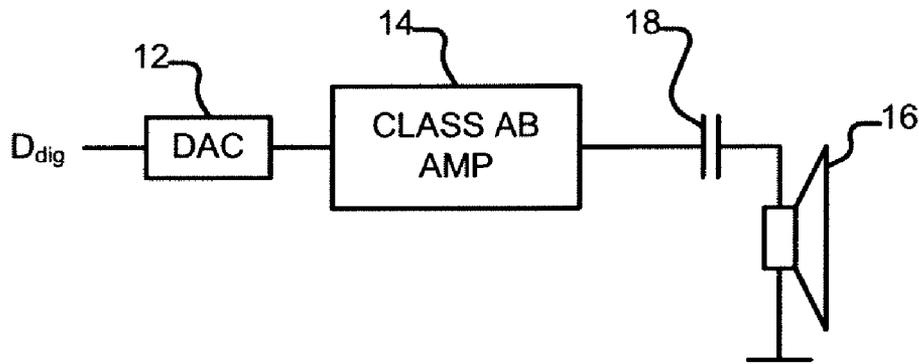
*Primary Examiner*—Khai M Nguyen

(57) **ABSTRACT**

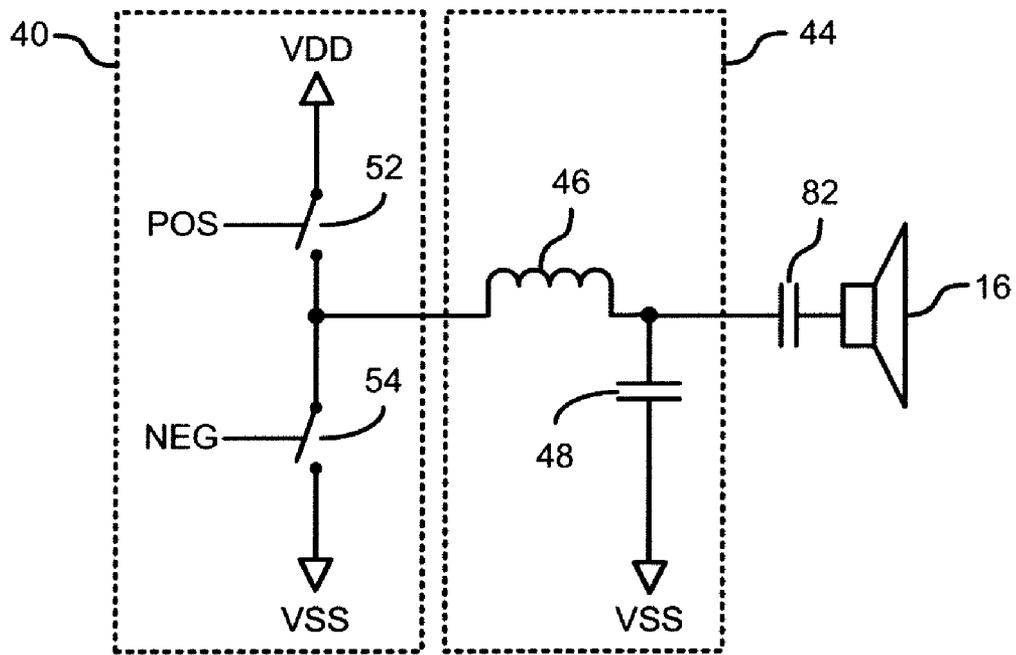
A noise shaper truncates the widths of pulses it supplies to a pulse-width modulator whose pulse-width modulated signals control a tri-level amplifier. The amplifier is filterless and DC free. Pulses that are narrower than a predefined minimum width are eliminated by the noise shaper. Other pulses are subjected to an algorithm that includes a multitude of ranges, each range defined by a minimum pulse width characterizing a lower bound of that range and a maximum pulse width characterizing an upper bound of that range. Associated with each range is a number of clock cycles defining the modified width of a pulses whose detected width falls within that range. To ensure that delays associated with a feedback loop disposed in the amplifier are accounted for, the reference voltage of a comparator tracks an output voltage of an integrator. Both the comparator and integrator are disposed in the loop.

**41 Claims, 18 Drawing Sheets**

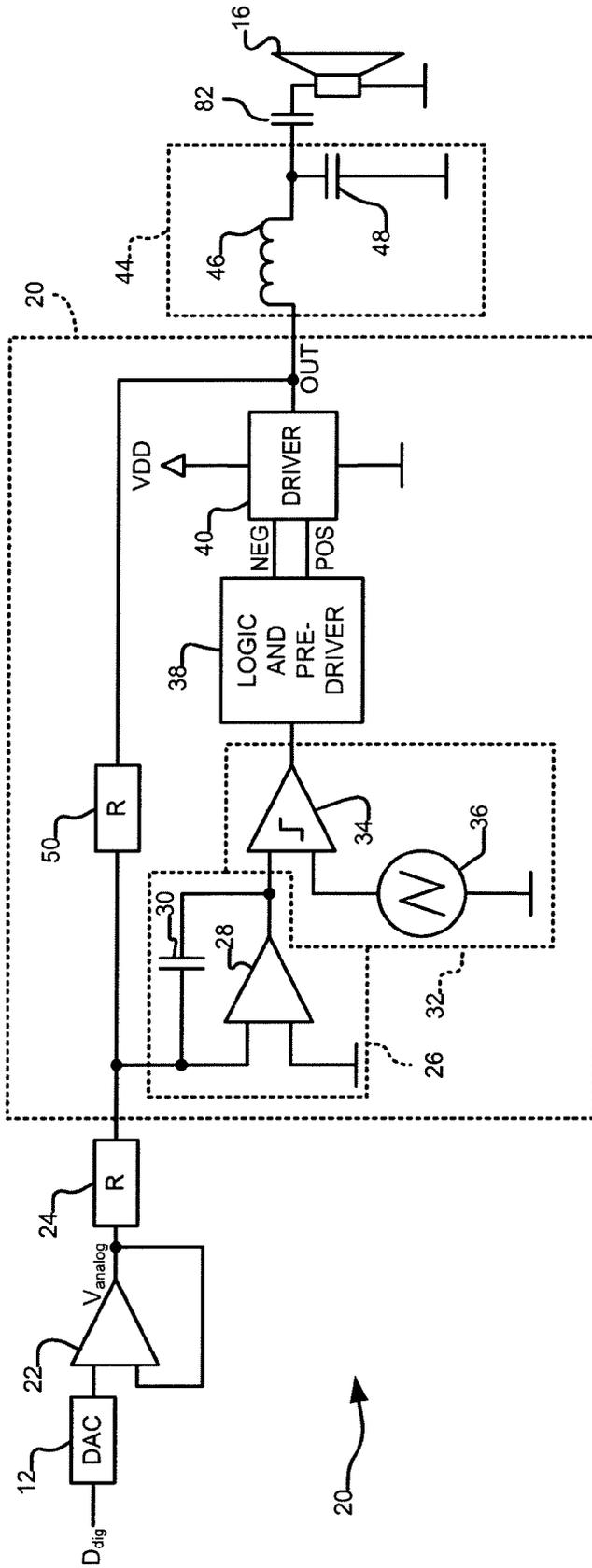




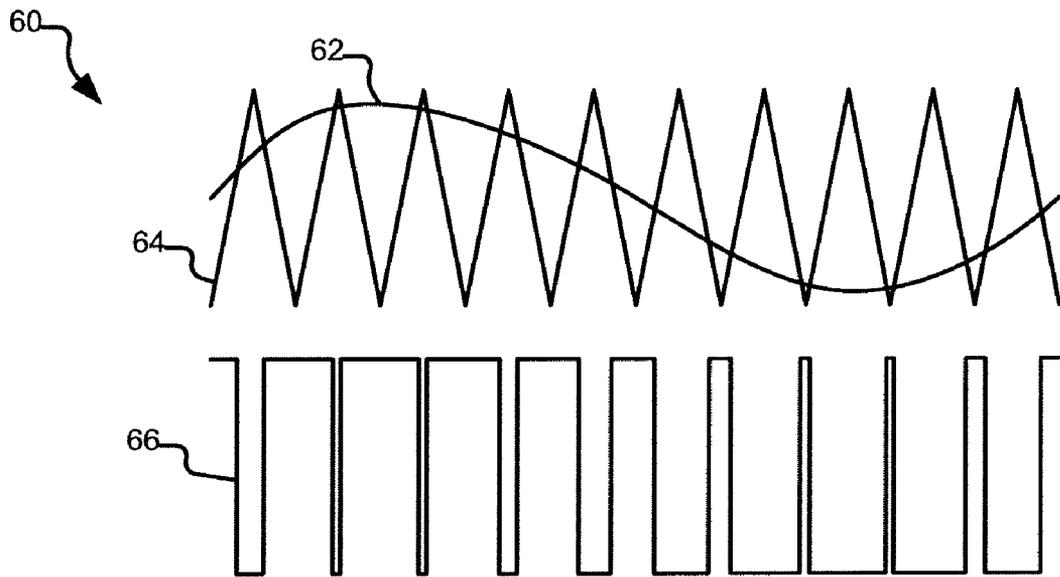
**FIG. 1**  
(PRIOR ART)



**FIG. 3**  
(PRIOR ART)



**FIG. 2**  
(PRIOR ART)



**FIG. 4**  
(PRIOR ART)



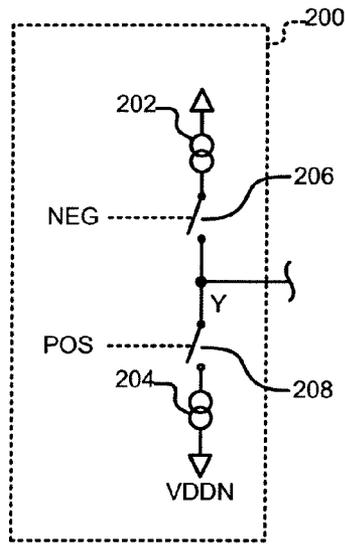


FIG. 6

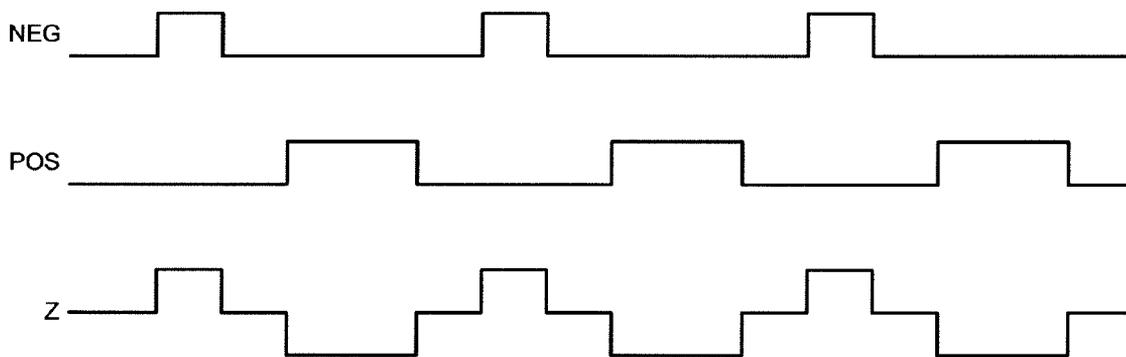


FIG. 7

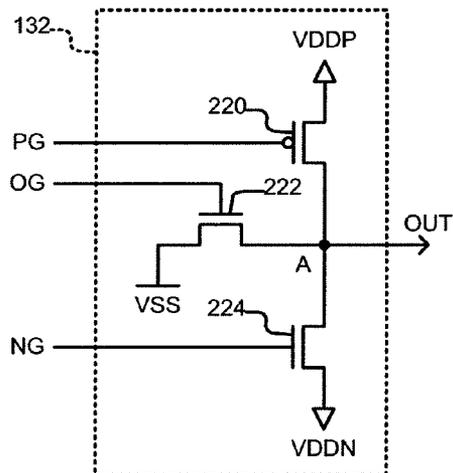


FIG. 8

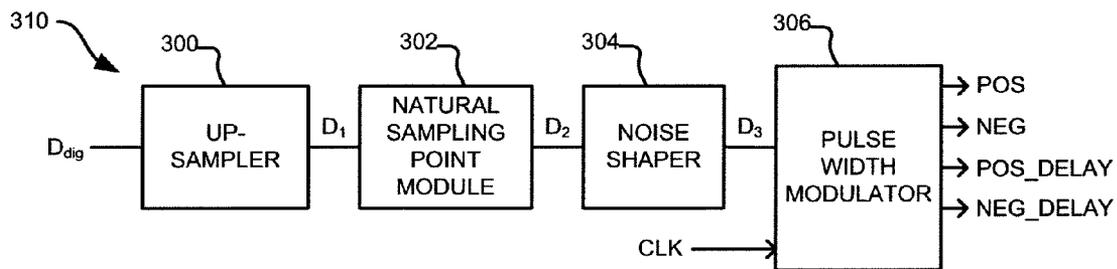


FIG. 9A

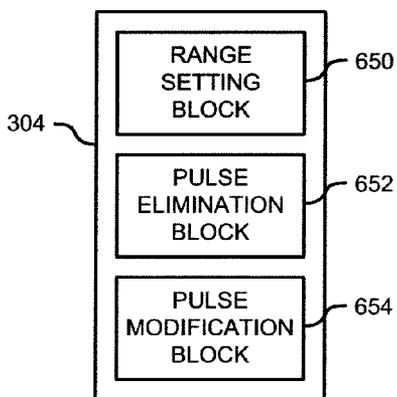
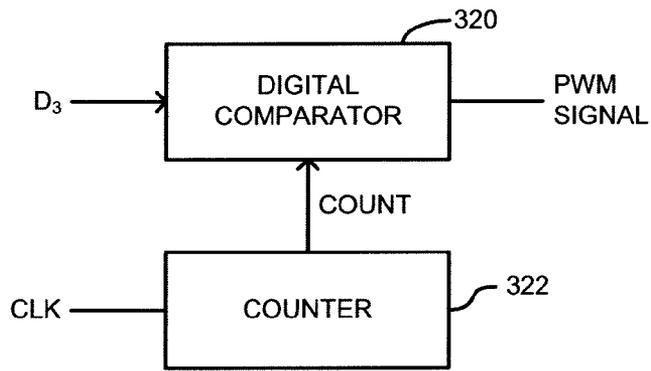
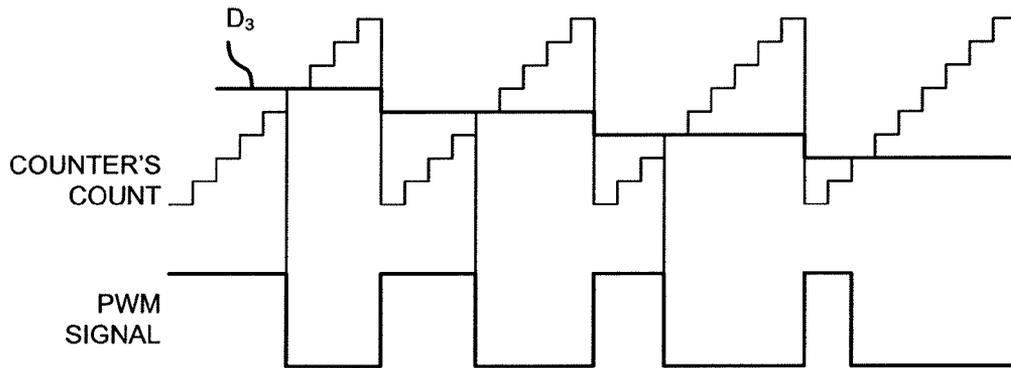


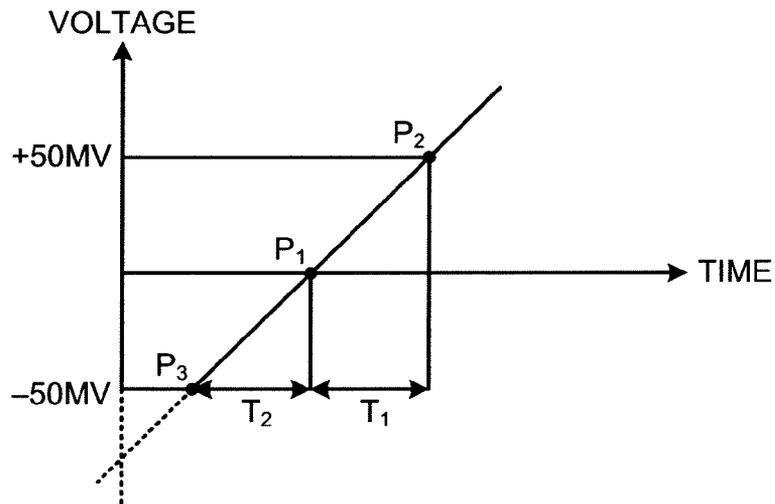
FIG. 9B



**FIG. 10**



**FIG. 11**



**FIG. 13**

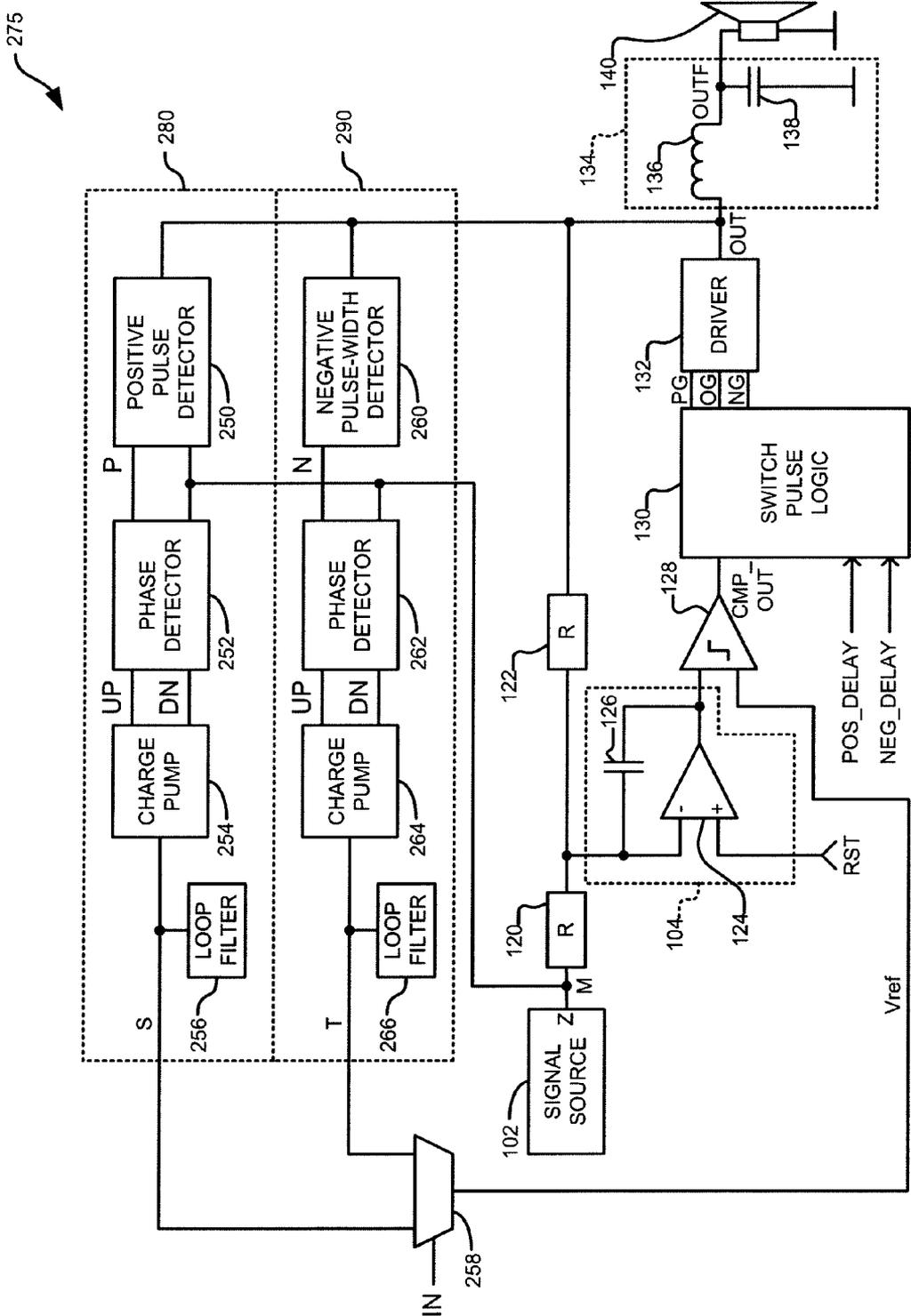


FIG. 12

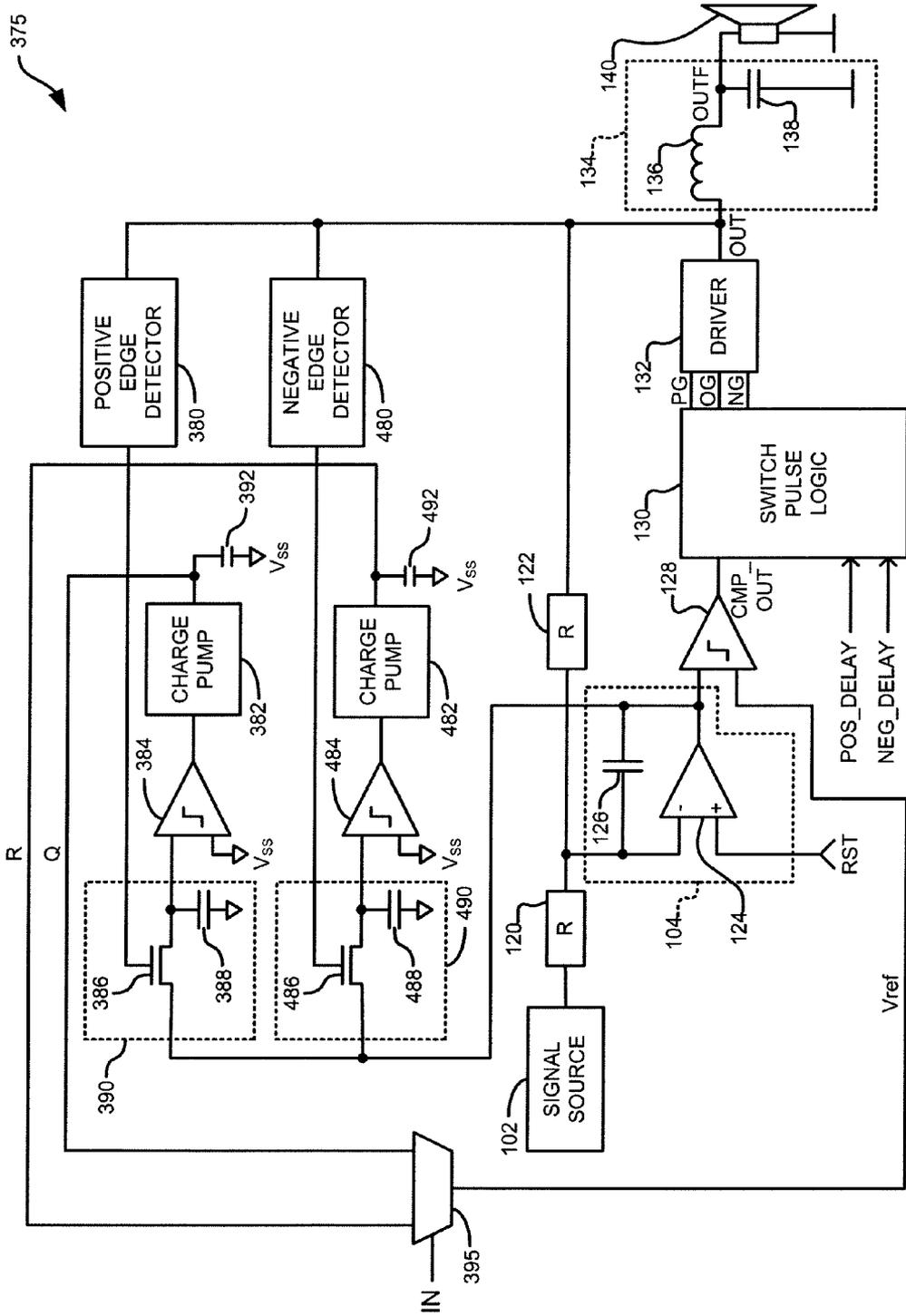


FIG. 14

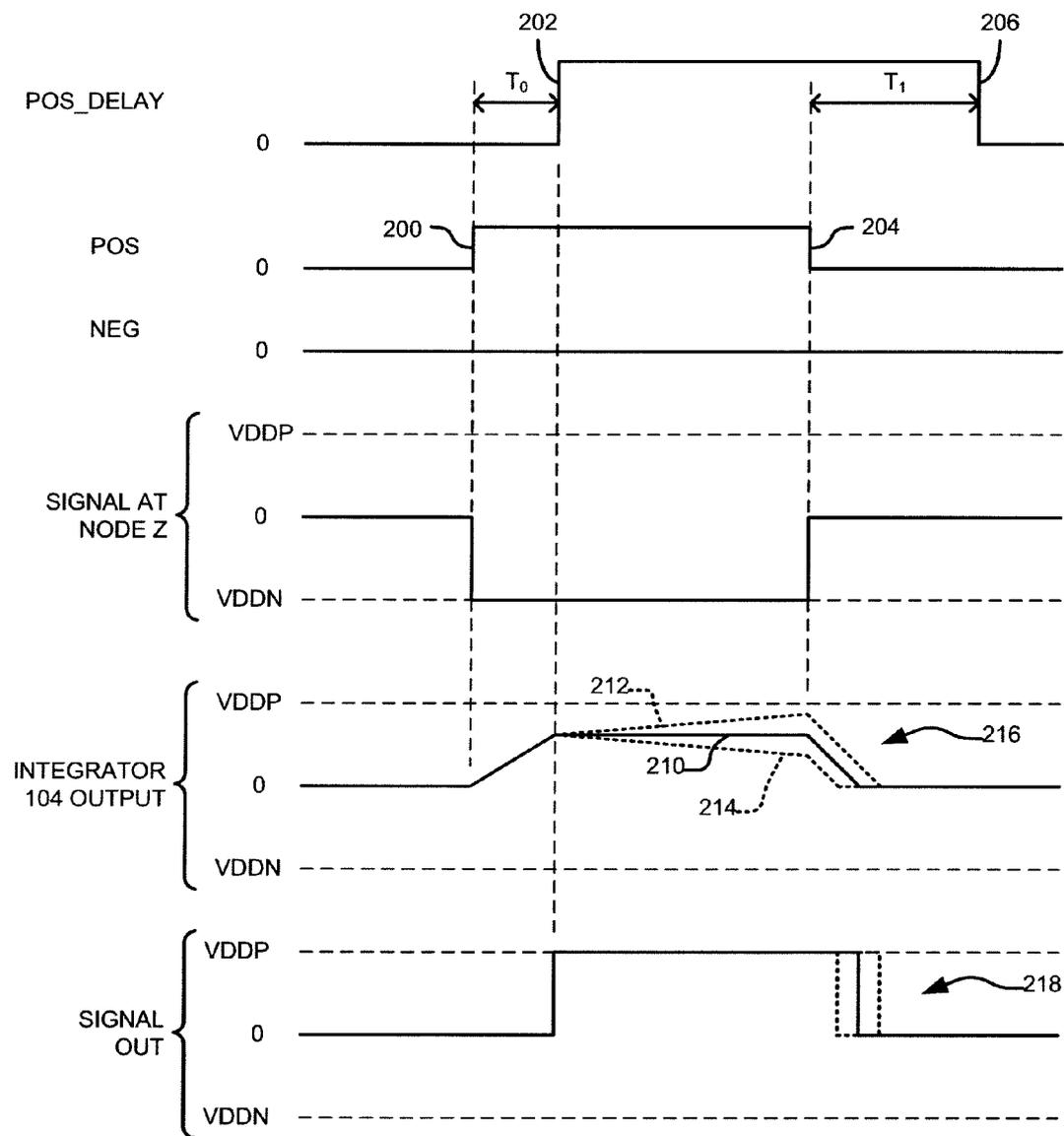


FIG. 15A

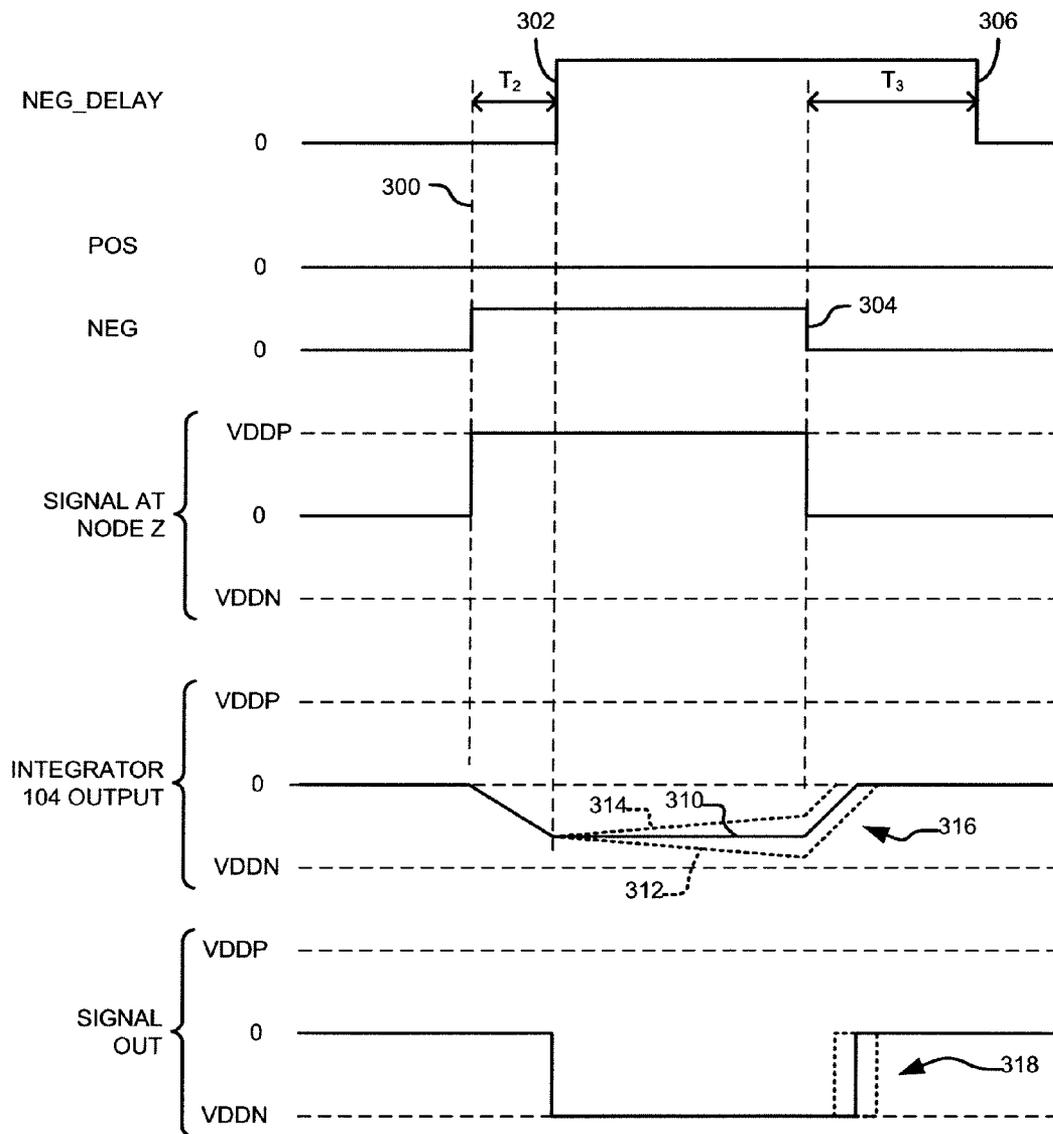
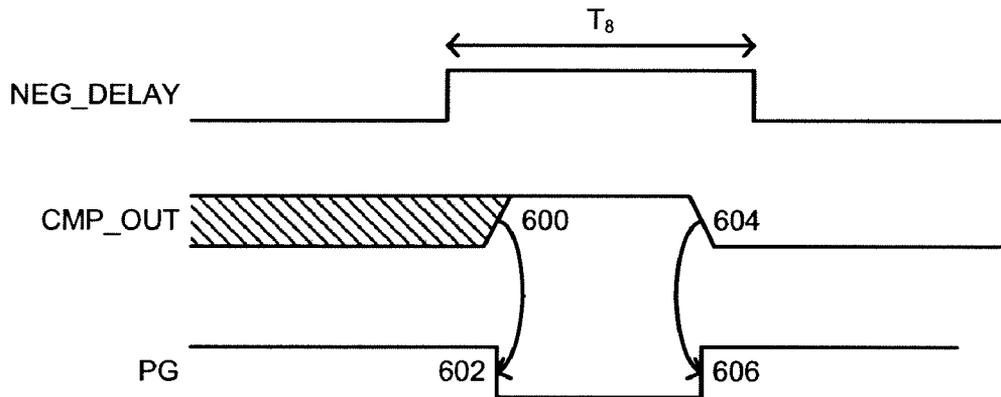
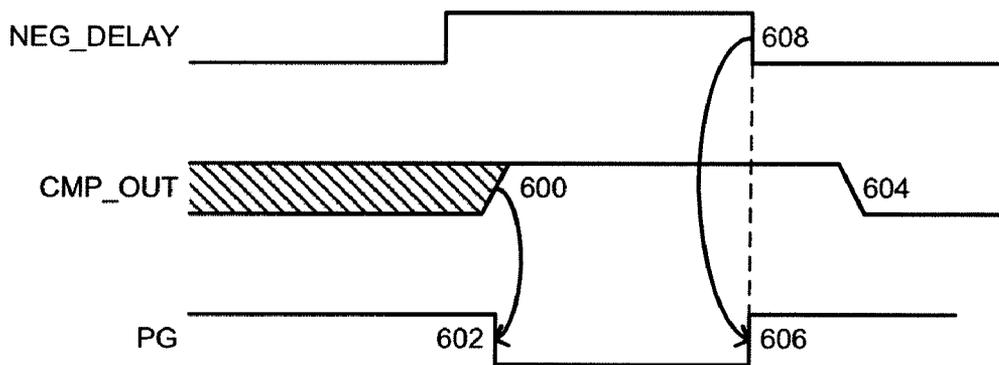


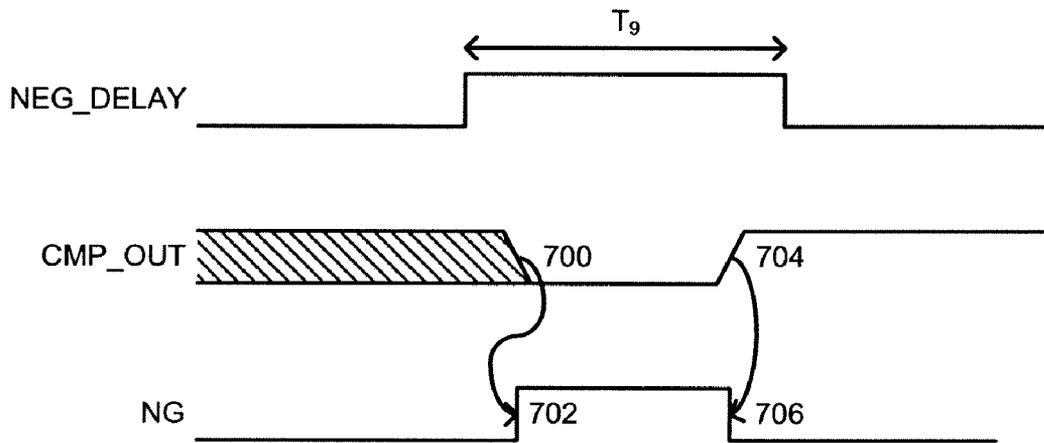
FIG. 15B



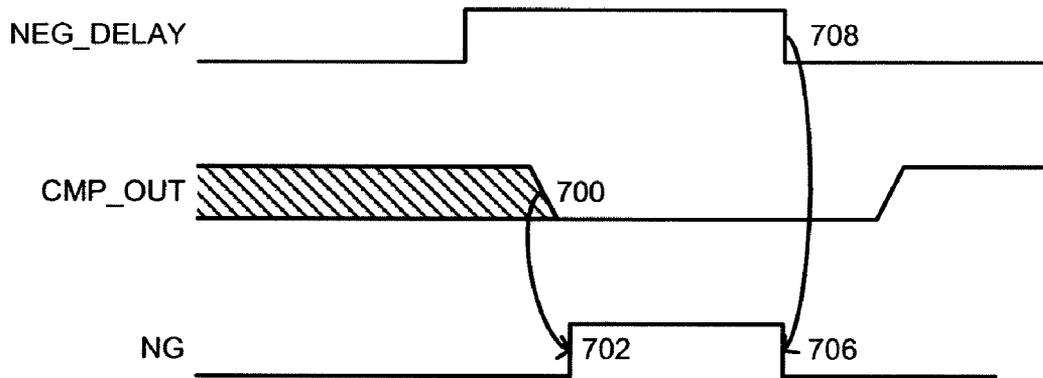
**FIG. 16A**



**FIG. 16B**



**FIG. 17A**



**FIG. 17B**

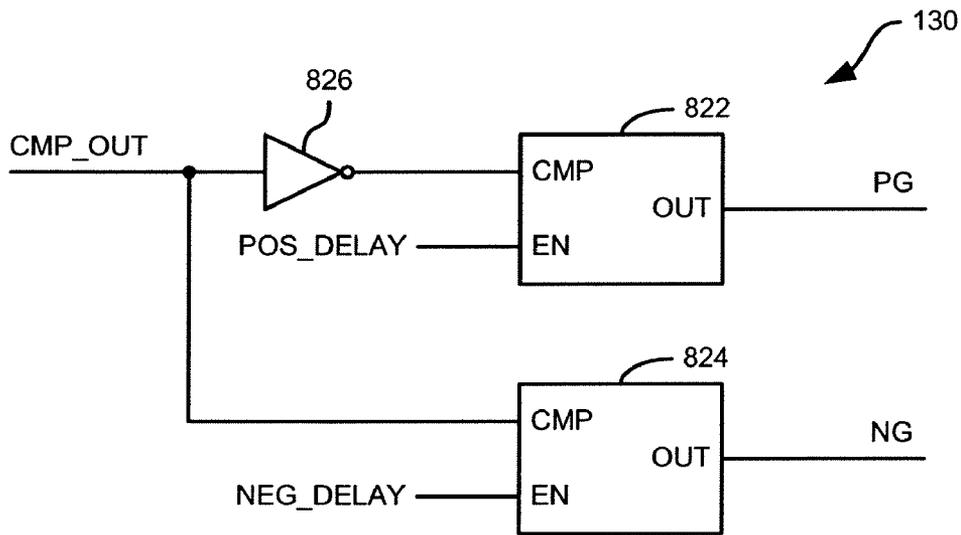


FIG. 18

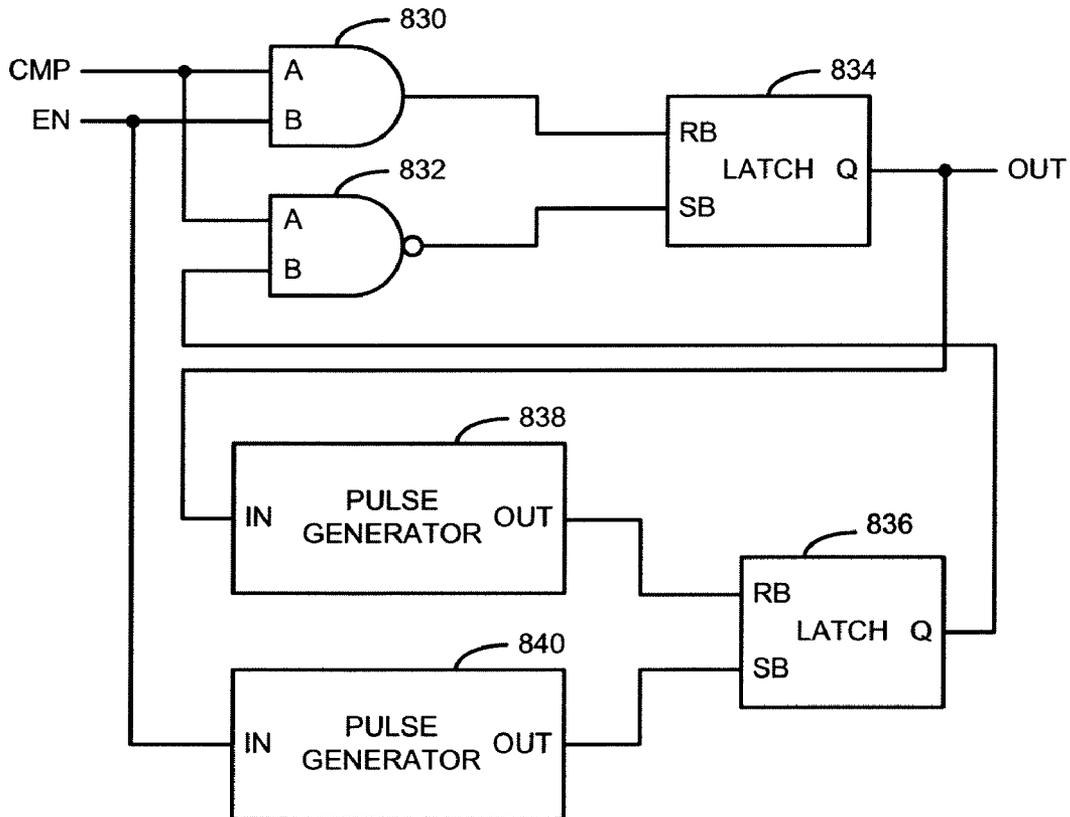


FIG. 19

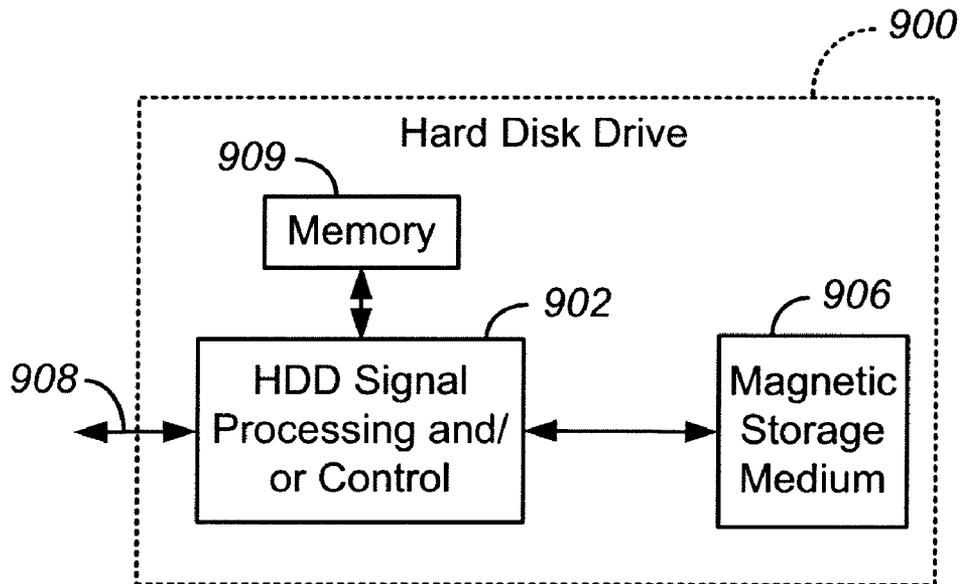


FIG. 20A

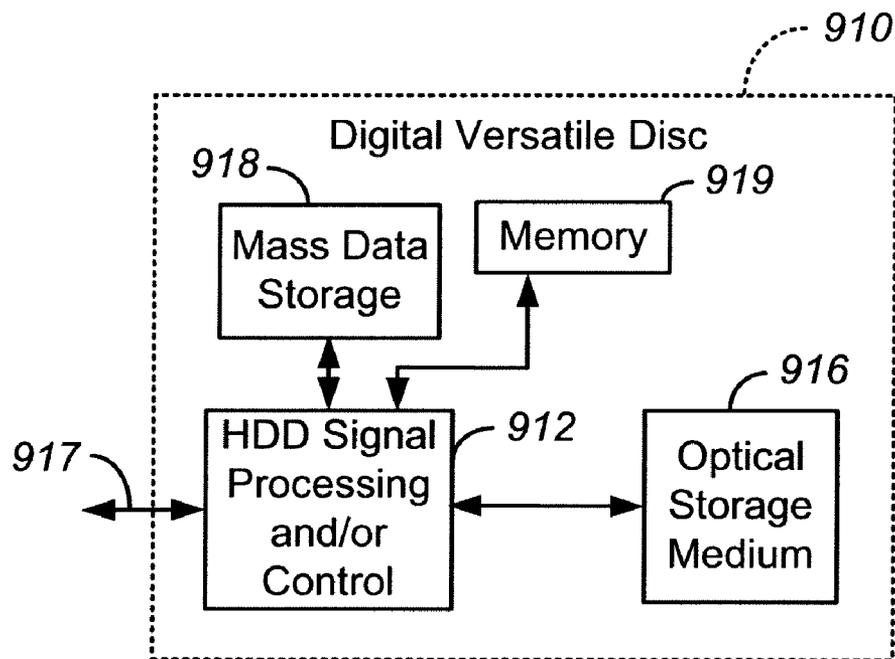


FIG. 20B

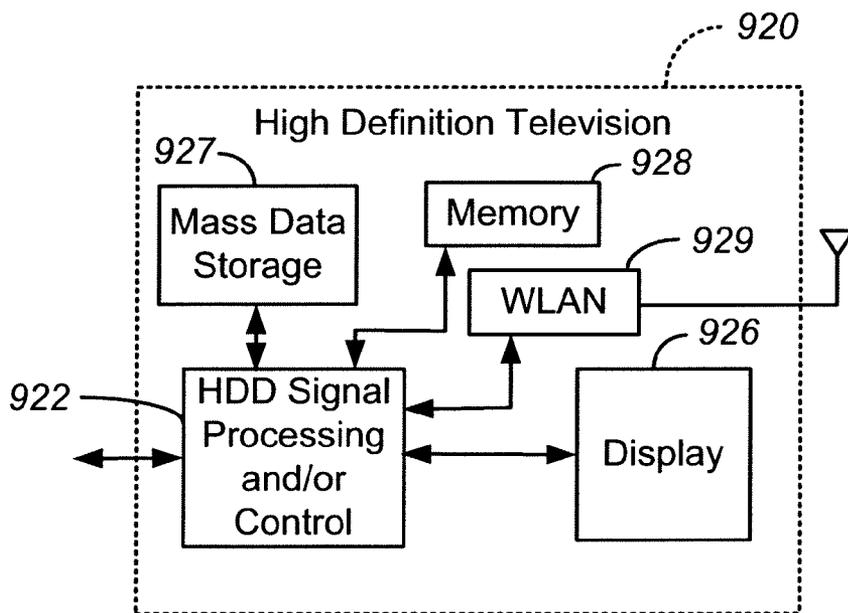


FIG. 20C

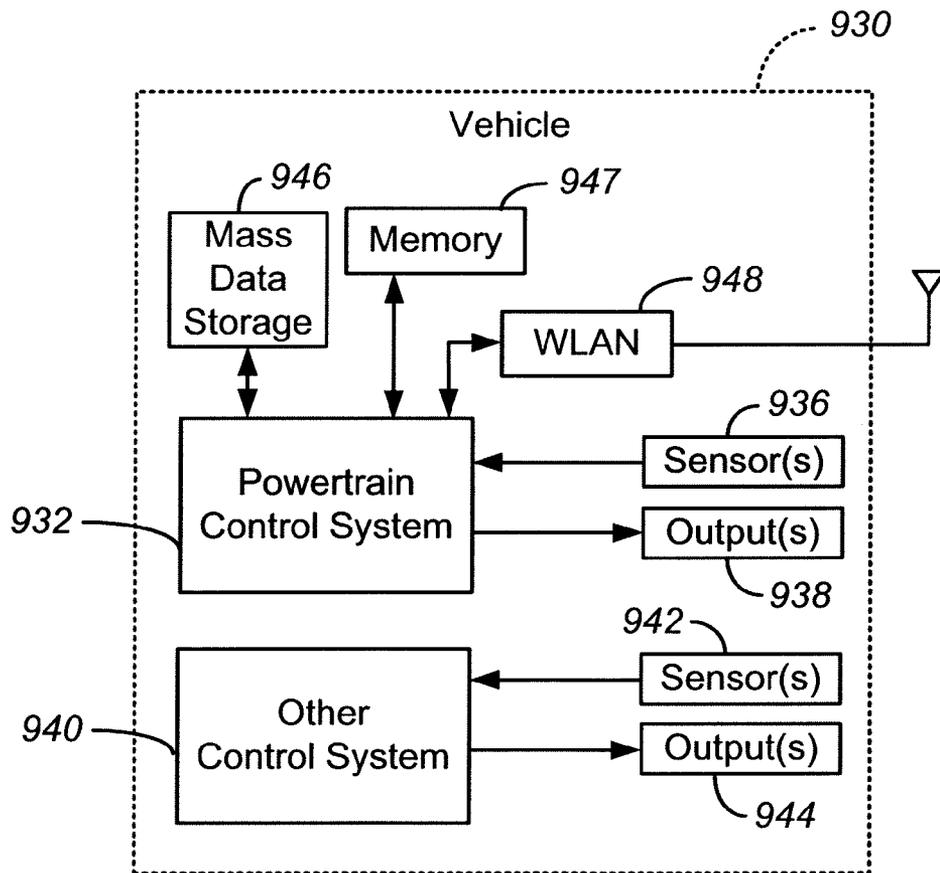


FIG. 20D

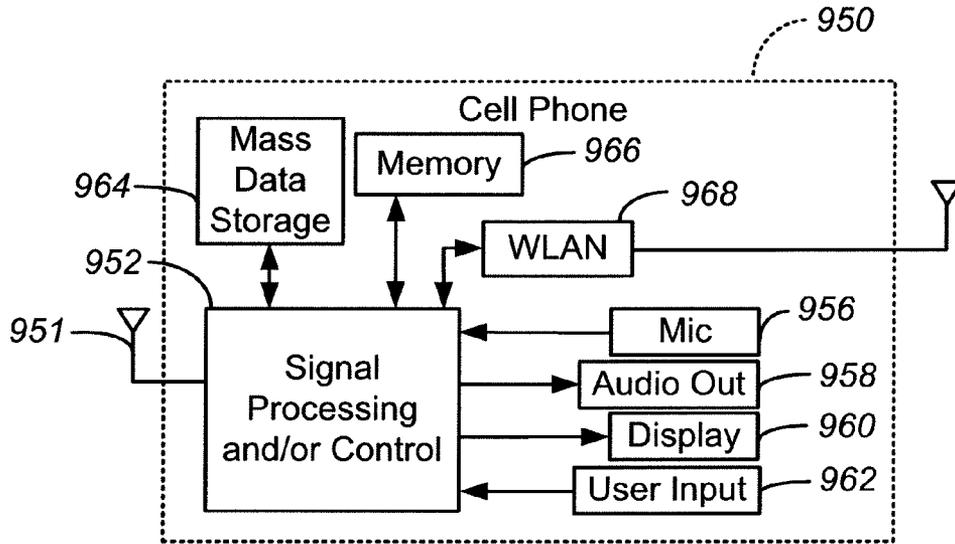


FIG. 20E

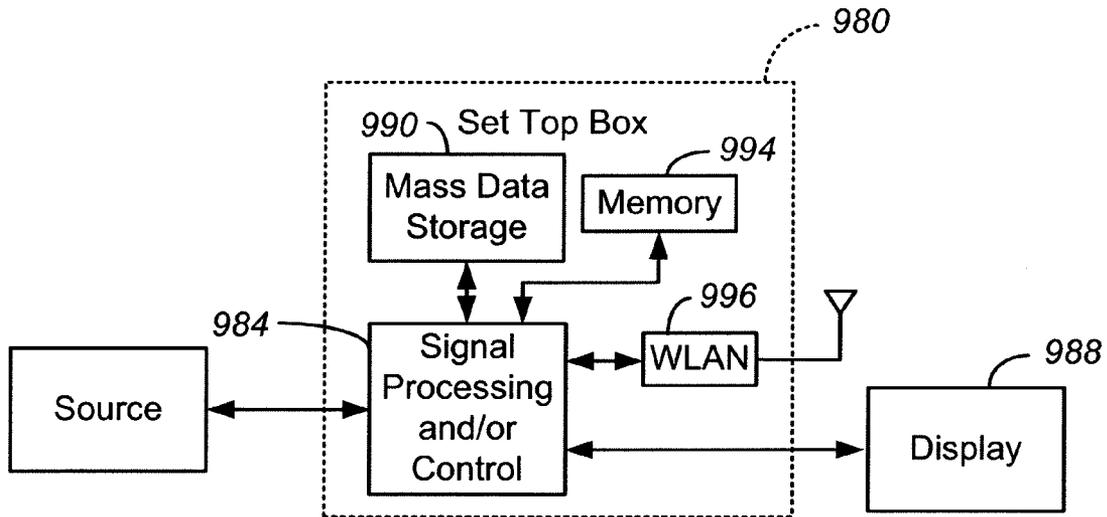


FIG. 20F

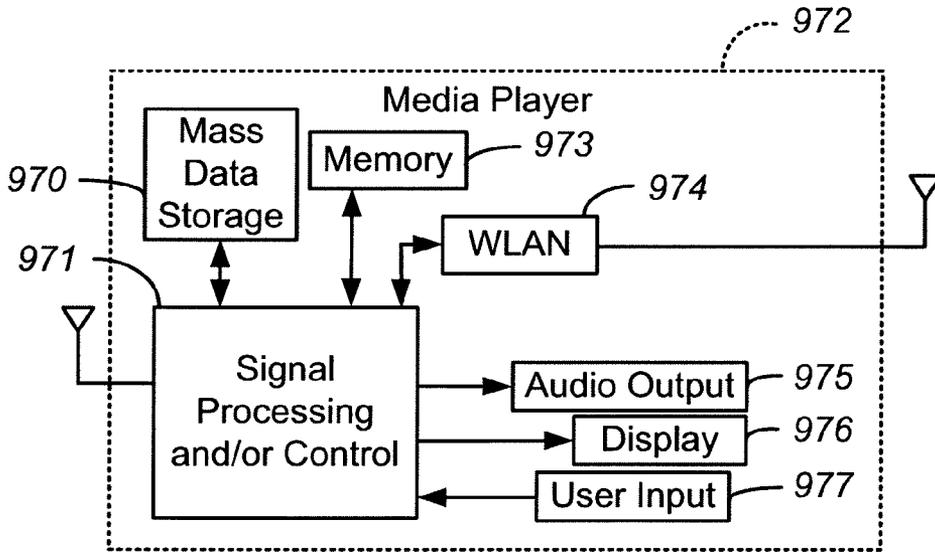


FIG. 20G

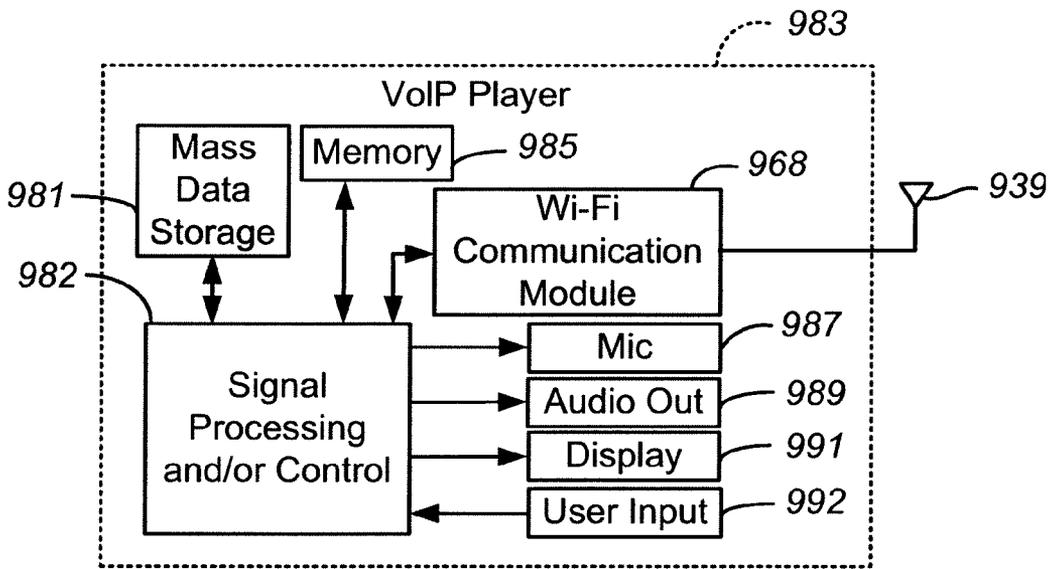


FIG. 20H

## TRUNCATION FOR THREE-LEVEL DIGITAL AMPLIFIER

### CROSS-REFERENCES TO RELATED APPLICATIONS

The present application claims benefit under 35 USC §119 (e) of U.S. provisional application No. 60/869,649, filed on Dec. 12, 2006, entitled "Methods of Truncation for 3-Level Digital Amplifier", the content of which is incorporated herein by reference in its entirety.

The present application is related to copending U.S. application Ser. No. 11/786,886, filed on Apr. 12, 2007, entitled "Power Efficient Amplifier", the content of which is incorporated herein by reference in its entirety.

### BACKGROUND OF THE INVENTION

The present application relates to amplifiers, and more particularly to amplifiers that have a reduced power consumption.

Audio amplifiers are well known and are used extensively to amplify audio signals. Designing an audio amplifier generally requires balancing two competing concerns. The first concern is fidelity which relates to the accuracy with which the audio amplifier reproduces the sounds contained in the audio signal. The second concern is power efficiency, which relates to the power consumption of the audio amplifier under various operating conditions.

FIG. 1 is a block diagram of an audio amplifier 10, as known in the prior art. Digital-to-analog converter (DAC) 12 converts the digital audio signal  $D_{dig}$  to an analog audio signal. The converted audio signal is applied to a class AB amplifier 14. The amplified audio signal is applied to speaker 16 via AC coupling capacitor 18. As is well known, amplifier 14 has a relatively low efficiency, thus rendering this amplifier undesirable for handheld portable devices which often have a limited battery life and/or internal cooling capacity.

FIG. 2 is a functional block diagram of a conventional analog class D audio amplifier 20. Class D audio amplifier 20 is generally more efficient than class AB amplifier 14 shown in FIG. 1. The fidelity of class D audio amplifier 20 can be comparable to that of Class AB audio amplifier 14 depending on variables such as switching frequency, thermal noise of resistors 24 and 50, noise of amplifier 28, noise and distortion of signal generator 36, noise of comparator 34, the loop delay and the blanking time of driver 40.

DAC 12 converts the digital audio signal  $D_{dig}$  to an analog audio signal which is subsequently amplified by operational amplifier (opamp) 22. Opamp 22 supplies the amplified audio signal to integrator 26 via resistive load 24. Integrator 26 includes an opamp 28 and a feedback capacitor 30. The output signal of integrator 26 is supplied to one of the input terminals of comparator 34. The other input terminal of comparator 34 receives a sawtooth or triangular waveform generated by sawtooth/triangular waveform generator 36. Comparator 34 and sawtooth/triangular waveform generator 36 together form a natural frequency sampling module 32 that generates a pulse-width modulated (PWM) signal. The frequency of the sawtooth/triangular waveform is usually at least 10 times higher than the maximum audible frequency included in the analog audio signal.

Logic and pre-driver 38 converts the received PWM signal into signals suitable for use by driver 40. In some embodiments, driver 40 is single-ended and includes a switch, such as a transistor and the like, that switches power to low-pass filter 44. Such a switch is opened and closed based on the

signals received from logic and pre-driver 38. The output signal of driver 40 is applied to low pass filter 44 and is also fed back to integrator 26 via resistor 50. Low-pass filter 44 removes the switching harmonics from the signal it receives via driver 40. Low-pass filter 44 is shown as including an inductor 46 and a capacitor 48. Signal OUT generated by driver 40 is delivered to speaker 16.

FIG. 3 is a block diagram of driver 40 coupled to low-pass filter 44 and speaker 16. Switches 52 and 54 are controlled by signals POS and NEG received from logic and pre-driver 38. To apply a positive pulse to speaker 16, switch 52 is turned on and switch 54 is turned off. To apply a negative pulse to speaker 16, switch 54 is turned on and switch 52 is turned off. Inductor 46 together with capacitor 48 form a low-pass filter that attenuate the high frequency signal components. An AC coupling capacitor 82 can be connected in series with speaker 16 to eliminate a direct current (DC) component of the signal applied to speaker 16.

FIG. 4 is a timing diagram of the input signals received and output signal generated by comparator 34. Sinusoidal signal 62 represents the input signal received from integrator 26. Sawtooth signal 64 represents the signal generated by waveform generator 36. Signal 66 represents the output signal of comparator 34 and is a pulse-width modulated (PWM) signal. The duty cycle of PWM signal 66 is determined by the amount of time that the magnitude of signal 62 is greater than the magnitude of signal 64. The period of PWM signal 66 is determined by the period of signal 64.

### BRIEF SUMMARY OF THE INVENTION

In accordance with one embodiment of the present invention, a method of truncating a pulse used to generate a tri-level amplified output signal includes, in part, establishing a multitude of ranges each defined by a minimum pulse width characterizing a lower bound of that range and a maximum pulse width characterizing an upper bound of that range, removing the pulse if the pulse width is narrower than a predefined value, and modifying the pulse width to a value characterized by truncating one of the lower or the upper bound of the range in which the pulse falls.

In one embodiment, the pulse width is modified to a value characterized by truncating down the upper bound of the range to its nearest integer value. In another embodiment, the pulse width is modified to a value characterized by truncating up the upper bound of the range to its nearest integer value. In yet another embodiment, the pulse width is modified to a value characterized by truncating up the lower bound of the range to its nearest integer value. In yet other embodiments, the pulse width is modified to a value characterized by truncating down the lower bound of the range to its nearest integer value.

In some embodiments, the difference between the predefined value resulting in elimination of the pulse and a truncated maximum bound of a succeeding range is greater than 1 clock cycle. In yet other embodiments, the pulse width is modified to an integer number of clock cycles defined by truncating to the integer number either the lower bound or the upper bound of the range in which the pulse falls.

In accordance with another embodiment of the present invention, a noise shaper includes, in part, circuitry adapted to remove a pulse if the pulse width is narrower than a predefined value, and circuitry adapted modify the pulse width to a value characterized by truncating one of the lower or the upper bounds of the range in which the pulse falls.

In one embodiment, the noise shaper further includes circuitry adapted to modify the pulse width to a value charac-

terized by truncating down the upper bound of the range to its nearest integer value. In another embodiment, the noise shaper further includes circuitry adapted to modify the pulse width to a value characterized by truncating up the lower bound of the range to its nearest integer value. In yet another embodiment, the noise shaper further includes circuitry adapted to modify the pulse width to a value characterized by truncating up the upper bound of the range to its nearest integer value. In yet other embodiments, the noise shaper further includes circuitry adapted to modify the pulse width to a value characterized by truncating down the minimum bound of the range to its nearest integer value.

In one embodiment, a difference between the predefined value and a truncated lower bound of a succeeding range is greater than 1 clock cycle. In yet another embodiment, the noise shaper further includes circuitry adapted to modify the pulse width to an integer number of clock cycles defined by truncating to the integer number either the lower bound or the upper bound of the range in which the pulse falls. In one embodiment, the noise shaper is disposed in an amplifier adapted to amplify audio/video signals.

In accordance with one embodiment of the present invention, a tri-level amplifier includes, in part, means for establishing a multitude of ranges each defined by a minimum pulse width characterizing a lower bound of that range and a maximum pulse width characterizing an upper bound of that range, means for removing the pulse if the pulse width is narrower than a predefined value, and means for modifying the pulse width to a value characterized by truncating one of the lower or the upper bound of the range in which the pulse falls.

In one embodiment, the pulse width is modified to a value characterized by truncating down the upper bound of the range to its nearest integer value. In another embodiment, the pulse width is modified to a value characterized by truncating up the upper bound of the range to its nearest integer value. In yet another embodiment, the pulse width is modified to a value characterized by truncating up the lower bound of the range to its nearest integer value. In yet other embodiments, the pulse width is modified to a value characterized by truncating down the lower bound of the range to its nearest integer value.

In some embodiments, the difference between the predefined value resulting in elimination of the pulse and a truncated maximum bound of a succeeding range is greater than 1 clock cycle. In yet other embodiments, the pulse width is modified to an integer number of clock cycles defined by truncating to the integer number either the lower bound or the upper bound of the range in which the pulse falls.

In accordance with another embodiment of the present invention, a circuit includes, in part, an integrator adapted to integrate a first signal in accordance with a feedback signal, a comparator responsive to an output of the integrator, a first pulse detector adapted to detect a pulse, a first phase detector responsive to an output of the first pulse detector and the first signal, and a first charge pump responsive to the first phase detector and adapted to generate a first reference voltage selectively applied to the comparator. In one embodiment, the pulse is a positive pulse.

In one embodiment, the circuit further includes a second pulse detector adapted to detect a negative pulse, a second phase detector responsive to an output of the second pulse detector and the first signal, a second charge pump responsive to the second phase detector and adapted to generate a second reference voltage, and a multiplexer adapted to select between the first and second reference voltages and apply the selected voltage to the comparator. In one embodiment the

circuit further includes a driver adapted to generate the positive pulse and the negative pulse at its output terminal. The circuit further includes a logic block adapted to control the driver in response to first and second pulse-width modulated signals. In one embodiment, the circuit is disposed in an amplifier adapted to amplify audio/video signals.

In accordance with another embodiment of the present invention, a method of canceling a loop delay includes, in part, integrating a first signal in accordance with a feedback signal, comparing the integrated signal to a first reference signal, detecting a pulse, detecting a difference between a phase of the pulse and a phase of the first signal, and varying the first reference signal in response to the detected phase difference. In one embodiment, the method further includes detecting a negative pulse, detecting a difference between a phase of the negative pulse and a phase of the first signal, and varying the first reference signal in response to the detected difference between the phase of the negative pulse and the phase of the first signal.

In one embodiment, the positive pulse varies between the ground potential and a positive supply voltage and the negative pulse varies between the ground potential and a negative supply voltage. In one embodiment, the method further includes generating the positive and negative pulses in response to first and second pulse-width modulated signals.

In accordance with another embodiment of the present invention, a tri-level amplifier includes, in part, means for integrating a first signal in accordance with a feedback signal, comparing the integrated signal to a first reference signal, detecting a pulse; means for detecting a difference between a phase of the pulse and a phase of the first signal, and means for varying the first reference signal in response to the detected phase difference. In one embodiment, the tri-level amplifier further includes means for detecting a negative pulse, means for detecting a difference between a phase of the negative pulse and a phase of the first signal, and means for varying the first reference signal in response to the detected difference between the phase of the negative pulse and the phase of the first signal.

In one embodiment, the positive pulse varies between the ground potential and a positive supply voltage and the negative pulse varies between the ground potential and a negative supply voltage. In one embodiment, the tri-level amplifier further includes means for generating the positive and negative pulses in response to first and second pulse-width modulated signals.

In accordance with one embodiment of the present invention, a circuit includes, in part, an integrator adapted to integrate a first signal in accordance with a feedback signal, a comparator responsive to an output of the integrator, a first edge detector adapted to detect a trailing edge of the feedback signal, a first sample-and-hold circuit responsive to an output of the integrator and the detected trailing edge, a second comparator responsive to an output of the first sample-and-hold circuit and to a supply potential, and a first charge pump responsive to an output of the second comparator and adapted to vary a first reference voltage selectively applied to the first comparator. In one embodiment, the first edge detector is a positive edge detector.

In one embodiment, the circuit further includes a negative edge detector adapted to detect a trailing edge of a negative pulse of the feedback signal, a second sample-and-hold circuit responsive to the output of the integrator, a third comparator responsive to an output of the second sample-and-hold circuit and to the supply potential, a second charge pump responsive an output of the third comparator and adapted to vary a second reference voltage, and a multiplexer adapted to

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select between the first and second reference voltages and apply the selected voltage to the first comparator. In one embodiment, the supply potential is the ground potential. In one embodiment, the circuit further includes a logic block adapted to supply first, second and third control signals in response to an output of the first comparator and further in response to first and second pulse width modulated signals. In one embodiment, the circuit further includes a driver adapted to generate the feedback signal in response to the first, second and third control signals. In one embodiment, the circuit is disposed in an amplifier adapted to amplify audio/video signals.

In accordance with another embodiment of the present invention, a method of minimizing effect of a loop delay on an output signal includes, in part, integrating a first signal in accordance with a feedback signal, comparing the integrated signal to a first reference signal, detecting a trailing edge of the feedback signal, sampling the integrated signal in response to the detected trailing edge; holding the sampled signal, comparing the held signal to a supply potential to generate a first compare signal, and varying the first reference signal in response to the first compare signal. In one embodiment, the trailing edge is a trailing edge of a positive pulse.

In some embodiments, the method further includes detecting a trailing edge of a negative pulse of the feedback signal, sampling the integrated signal in response to the detected trailing edge of the negative pulse, holding the signal sampled in response to the detected trailing edge of the negative pulse, comparing the signal held in response to the detected trailing edge of the negative pulse to the supply potential to generate a compare signal, and varying the first reference signal in response to the second compare signal.

In one embodiment, the supply potential is the ground potential. In another embodiment, the positive pulse varies between the ground potential and the positive supply voltage, and the negative pulse varies between the ground potential and the negative supply voltage. In one embodiment the method further includes generating the feedback signal in response to first and second pulse-width modulated signals.

In accordance with another embodiment of the present invention, a tri-level amplifier includes, in part, means for integrating a first signal in accordance with a feedback signal, means for comparing the integrated signal to a first reference signal, means for detecting a trailing edge of the feedback signal, means for sampling the integrated signal in response to the detected trailing edge, means for holding the sampled signal; means for comparing the held signal to a supply potential to generate a first compare signal, and means for varying the first reference signal in response to the first compare signal. In one embodiment, the trailing edge is a trailing edge of a positive pulse.

In some embodiments, the tri-level amplifier further includes means for detecting a trailing edge of a negative pulse of the feedback signal, means for sampling the integrated signal in response to the detected trailing edge of the negative pulse, means for holding the signal sampled in response to the detected trailing edge of the negative pulse, means for comparing the signal held in response to the detected trailing edge of the negative pulse to the supply potential to generate a compare signal, and means for varying the first reference signal in response to the second compare signal.

In one embodiment, the supply potential is the ground potential. In another embodiment, the positive pulse varies between the ground potential and the positive supply voltage, and the negative pulse varies between the ground potential and the negative supply voltage. In one embodiment the tri-

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level amplifier further includes means for generating the feedback signal in response to first and second pulse-width modulated signals.

#### BRIEF DESCRIPTION OF THE DRAWINGS

FIG. 1 is a functional block diagram of a Class AB amplifier, as known in the prior art:

FIG. 2 is a functional block diagram of a Class D amplifier, as known in the prior art.

FIG. 3 is a schematic diagram of the driver disposed in the amplifier of FIG. 2, as known in the prior art.

FIG. 4 is timing diagram of the input/output signals of the comparator disposed in the amplifier of FIG. 2, as known in the prior art.

FIG. 5 is a functional block diagram of an amplifier, in accordance with one embodiment of the present invention.

FIG. 6 is a functional block diagram of an exemplary signal generator adapted to supply a signal to the amplifier of FIG. 5, in accordance with one embodiment of the present invention.

FIG. 7 is a timing diagram of a number of signals associated with the amplifier of FIG. 5, in accordance with one embodiment of the present invention.

FIG. 8 is a schematic diagram of an exemplary driver disposed in the amplifier of FIG. 5, in accordance with one embodiment of the present invention.

FIG. 9A is a diagram of exemplary logic blocks used to control the amplifier of FIG. 5, in accordance with one embodiment of the present invention.

FIG. 9B is an exemplary logic block diagram of the noise shaper of FIG. 9A, in accordance with one embodiment of the present invention.

FIG. 10 is a block diagram of a number of components adapted to generate a pulse-width modulated signal.

FIG. 11 is an exemplary timing diagram of the signals associated with the pulse-width modulator of FIG. 10.

FIG. 12 is a functional block diagram of an amplifier, in accordance with another embodiment of the present invention.

FIG. 13 shows an exemplary output voltage of the integrators disposed in the amplifiers of FIGS. 5 and 12.

FIG. 14 is a functional block diagram of an amplifier, in accordance with another embodiment of the present invention.

FIGS. 15A-15B are exemplary timing diagrams of a number of signals associated with the amplifier of FIG. 5.

FIGS. 16A-16B are exemplary timing diagrams of a number of signals associated with the switch pulse logic of the amplifier of FIG. 5.

FIGS. 17A-17B are exemplary timing diagrams of a number of signals associated with the switch pulse logic of the amplifier of FIG. 5.

FIG. 18 is a logic block diagram of the switch pulse logic disposed in the amplifier of FIG. 5, in accordance with one embodiment of the present invention.

FIG. 19 shows a number of logic blocks disposed in the switch pulse logic of FIG. 16, in accordance with one embodiment of the present invention.

FIGS. 20A-20H show various devices in which the present invention may be embodied.

#### DETAILED DESCRIPTION OF THE INVENTION

An amplifier, in accordance with one embodiment of the present invention, generates a tri-level output signal in response to a digital input signal. The amplifier is filterless and DC free. The amplifier includes, in part, a signal genera-

tor, an integrator, a comparator, a switch pulse logic, a driver, and a control block adapted to supply a multitude of pulse-width modulated (PWM) signals to the switch pulse logic in response to the received digital input signal. A pair of the PWM signals are applied to the signal generator which in response supplies a signal to the integrator. The integrator's output signal is compared to a reference signal by the comparator. The switch pulse logic receives the output of the comparator and a pair of delayed PWM signals, and in response generates a multitude of driver signals applied to the driver. In response, the driver supplies an output signal that is adapted to vary between first, second and third voltages. In one embodiment, the third voltage may be the ground voltage.

FIG. 5 is a block diagram of an amplifier 100 that provides a tri-level output signal, in accordance with one embodiment of the present invention. The following description is provided with reference to using amplifier 100 in audio applications. It is understood, however, that amplifier 100 may also be used in many other applications, such as in processing video signals, etc.

Signal generator 102, also referred to herein as signal source, is shown as including voltage dividers 175, 180, capacitors 160, 162, operational amplifiers 164, 166 and switches 168 and 170. In response to signals POS and NEG applied to switches 168 and 170 from a control block, current is either sourced to or drawn from output node Z of signal source 102. In some embodiments, switches 168 and 170 may be transistor switches.

Voltage divider 175 includes resistive loads 112 and 114 that are disposed between positive supply voltage VDDP and a reference voltage Vss. Reference voltage Vss may be the ground voltage. The inverting input of operational amplifier (op-amp) 164 is coupled to its output. Op-amp 164 generates a voltage at node W that is maintained substantially at the same potential as that at node V. Capacitor 162 attenuates the high frequency components of noise present at node V. Voltage divider 180 includes resistive loads 116 and 118 that are disposed between negative supply voltage VDDN and the reference voltage Vss. The inverting input of op-amp 166 is coupled to its output. Op-amp 166 generates a voltage at node X that is maintained substantially at the same potential as that at node U. Capacitor 160 attenuates the high frequency components of any noise present at node U. In some embodiments the feedback path between the input and output terminals of op-amps 164 and 166 may include a resistor or a compensation network. When signal NEG is asserted, signal POS is deasserted, accordingly the voltage at node W is supplied to node Z. Conversely, when signal POS is asserted, signal NEG is deasserted, accordingly the voltage at node X is supplied to node Z. The signal supplied to node Z by signal source 102 has a substantially attenuated noise component.

FIG. 6 is a schematic diagram of another exemplary embodiment 200 of a signal source that may be used in place of signal source 102 and resistor 120 of FIG. 5. Signal source 200 is shown as including current sources 202, 204 and switches 206, 208. Switch 206 that receives signal NEG is disposed between current source 202 and node Y, shown in FIG. 5. Similarly, switch 208 that receives signal POS is disposed between current source 204 and node Y. When signal NEG is asserted, the current supplied by current source 202 flows to node Y. When signal POS is asserted, the current supplied by current source 204 flows to node Y.

FIG. 7 is a timing diagram of voltage signals POS, NEG as well as the voltage signal supplied to node Z. Referring to FIGS. 6 and 8 concurrently, when signal POS is at a high level, the voltage at node Z corresponds to the voltage at node X, and when signal NEG is at a high level, the voltage at node

Z corresponds to the voltage at node W. When signals POS and NEG are both at zero levels, signal Z is also at the reference voltage Vss, e.g., zero volts. In one embodiment, supply voltages VDDP and VDDN are respectively at +1.5 volts and -1.5 volts. Accordingly, in such embodiments, the voltage at node Z varies between zero volts and a fraction of  $\pm 1.5$  volts.

Referring to FIG. 5, the signal present at node Z is delivered to an input terminal of op-amp 124 via resistor 120. The other input terminal of op-amp 124 is coupled to a reference voltage  $V_{ss}$  that may or may not be the ground potential. Feedback capacitor 126 disposed between input and output terminals of op-amp 124 is discharged when signal RST applied to switch 190 is asserted. Op-amp 124 together with capacitor 126 form an integrator 104. Op-amp 124 may be powered by supply voltages VDDP and VDDN.

The output signal of integrator 104 is applied to a first terminal of comparator 128 whose other terminal receives reference voltage Vref. In some embodiments, voltage Vref may be the ground potential. If the voltage supplied by integrator 104 is greater than reference voltage Vref, the output signal of comparator 128 is set to a high level, and if the voltage supplied by integrator 104 is smaller than reference voltage Vref, the output signal of comparator 128 is set to a low level. Comparator 128 may be powered by supply voltages VDDP and VDDN.

Switch pulse logic 130 receives the output signal of comparator 128 as well as the pulse-width modulated (PWM) signals POS\_DELAY and NEG\_DELAY, and in response generates control signals NG, PG and OG that are applied to driver 132. Driver 132 generates a tri-level output signal OUT in response to control signals NG, PG and OG that it receives. In one embodiment, such as that shown in FIG. 5, signal OUT is applied to a low pass filter 134, shown as including an inductor 136 and a capacitor 138. It is understood, however, that other low-pass filters may be used. The filtered output signal OUTF is applied to speaker 140. Signal OUT is fed back to the inverting terminal of op-amp 124 via resistive load 122. In some embodiments (not shown), an inductor and a capacitor internal to speaker 140 is used to filter out signal OUT. Accordingly in such embodiments, signal OUT is applied to directly to speaker 140 or a headphone. Amplifier 100 is filterless. Furthermore, since the DC level of the voltage signal supplied by amplifier 100 does not need to be shifted with respect to a reference DC voltage, amplifier 100 is DC free. In other words, to deliver, for example, 1.5 volts, the output voltage of amplifier 100 is set to 1.5 volts and not to another DC voltage level.

FIG. 8 is a transistor schematic diagram of an exemplary embodiment of driver 132. Driver 132 is shown as including transistors 220, 222 and 224 that respectively receive signals PG, OG and NG at their gate terminals. When signal PG is at a low level, signals NG and OG are at a high level. Accordingly, PMOS transistor 220 is on and NMOS transistor 222 and 224 are off. Therefore, node A is pulled to the positive supply voltage VDDP. When signals PG and NG are at a high level, signal OG is at a low level. Accordingly, transistor 224 is on and transistors 220 and 222 are off. Therefore, node A is pulled to the negative supply voltage VDDN. When signal PG is at a high level and signal NG is at a low level, signal OG is at a high level. Therefore, node A is pulled to the supply voltage Vss. Consequently, driver 132 is adapted to vary the voltage at node A between supply voltages VDDP, VDDN and Vss. In other words, driver module 132 generates a tri-level output signal by only turning on one of the transistors 220, 222 and 224 at any given time.

The feedback loop from the output of driver 132 to the input of op-amp 124 provides a number of advantages. First, the feedback loop attempts to compensate for the finite switch resistance at output node A of driver 132. Ideally, the switch resistance seen by node A should be as small as possible, e.g., 0 ohms, to enable node A to be pulled to supply voltages VDDP, VDDN and Vss as closely as possible. Second, the feedback loop compensates for the voltage overshoot and undershoot at node A during blanking times, as described further below. To prevent current flow from the positive voltage supply to the negative voltage supply, any of the transistors 220, 222 and 224 that is on during one cycle is first turned off before another one of these transistors is turned on during the next cycle. The time period between turning off of a transistor and turning on of another transistor in an output driver is referred to as the blanking time.

When, for example, transistor 220 is turned off, the current flow through the inductive load 136 seen by node A does not change instantaneously (see FIG. 5). This causes the voltage at node A to increase, causing a supply overshoot. The overshoot causes the source-to-bulk diode of transistor 220 to turn on. Once this diode is turned on, the voltage at node A is clamped at one PN junction diode above the VDDP voltage. Similarly, a voltage undershoot can occur during switching of transistor 224. As described above, the feedback loop compensates for the supply voltage overshoot and undershoot at output node A. Third, the feedback loop helps achieve power supply rejection. In other words, the feedback loop rejects variations in the voltage at node A that may arise due to variations in supply voltages VDDP and VDDN.

FIG. 9A shows various components of a control block 310 adapted to control amplifier 100 of FIG. 5. In response to the digital audio signal  $D_{dig}$ , control block 310 generates signals POS, NEG, POS\_DELAY and NEG\_DELAY that are applied to amplifier 100 shown in FIG. 5. In some embodiments, the audio signal  $D_{dig}$  is compliant with an I<sup>2</sup>S (Inter-IC Sound) bus specification published by Philips Corporation; this specification is hereby incorporated by reference in its entirety. In some embodiments, the audio signal  $D_{dig}$  may include audio samples between about 8 KHz and 192 KHz.

Up-sampler 300 upconverts the sampling rate of signal  $D_{dig}$ . In one embodiment, signal  $D_{dig}$  has a sampling rate of, for example, 48 KHz and signal  $D_1$  supplied by up-sampler 300 has a sampling rate of, for example, 384 KHz. Natural sampling point module 302 identifies the cross sections of a sawtooth/triangular signal and the up-sampled signal  $D_1$  to generate a digitized word  $D_2$ . In one embodiment,  $D_2$  is, for example, a 20-bit word and thus a clock signal of, for example, 400 GHz clock would be required to process the, e.g., 20-bit word  $D_2$ .

To dispense the need for such a high frequency clock, e.g., a 400 GHz clock, signal  $D_2$  is applied to noise shaper 304 which enables pulse-width modulator 306 to use a slower clock, e.g., a 480 MHz clock. Noise shaper 304 performs a bit-reduction algorithm and spreads the quantization error around so that more of the error is focused on higher inaudible frequencies. The result is that where human ear is most sensitive, the quantization error is reduced, and where human ear is less sensitive, the noise is greater. In one embodiment noise shaper 304, reduces the number of bits in signal  $D_2$  from, e.g., 20 bits to generate signal  $D_3$  that has, e.g., 8-bits. Some embodiments of the present invention do not include a noise shaper. In such embodiments, the output signal  $D_2$  of natural sampling module 302 is applied directly to pulse-width modulator 306.

Pulse width modulator 306 includes one or more counters and one or more digital comparators that generate pulse-width modulated signals POS, NEG, POS\_DELAY and NEG\_DELAY. FIG. 10 is a block diagram of a counter 322 coupled to a digital comparator 320 adapted to generate a

pulse-width modulated signal. Counter 322 increments its count with each transition of clock signal CLK. The count of counter 322 establishes the comparison value of comparator 320. When signal  $D_3$  is greater than the counter's count, comparator 320 generates a logic high signal, and when signal  $D_3$  is smaller than the counter's count, comparator 320 generates a logic low signal. Accordingly, the width of the pulses generated by comparator 320 is modulated in accordance with signal  $D_3$  and the counter's count. FIG. 11 is an exemplary timing diagram of the counter 320's count, signal  $D_3$  and the resulting pulse-width modulated signal.

Signal POS\_DELAY has a phase delay with respect to signal POS. In some embodiment, the delay between phases of signals POS and POS\_DELAY is fixed. In other embodiments, the delay between phases of signals POS and POS\_DELAY is programmable and thus may be varied. Similarly, signal NEG\_DELAY has a phase delay with respect to signal NEG. In some embodiment, the delay between phases of signals NEG and NEG\_DELAY is fixed. In other embodiments, the delay between phases of signals NEG and NEG\_DELAY is programmable and thus may be varied.

A pulse with a width that is less than a threshold value may not be properly transferred. Therefore, in accordance with the present invention, noise shaper 304 performs pulse width truncation to ensure that pulses reaching driver 132 have the required minimum width to bring about the expected results. In accordance with this truncation algorithm, a number of pulse width ranges are defined. Associated with each range is a multitude of clock cycles defining the modified width of a pulse falling within that range. In one embodiment, the multitude of clock cycles associated with each range is defined by truncating down the upper bound of that range to its nearest integer value. In another embodiment, the multitude of clock cycles associated with each range is defined by truncating up the lower bound of that range to its nearest integer value. For example, if the lower and upper bounds of a range are respectively 8.5 and 9.49 clock cycles, any pulse having a width falling within this range is modified to have a pulse width of 9 clock cycles, obtained by truncating down 9.49 to its nearest integer or truncating 8.49 up to its nearest integer value.

In general, a pulse having a width that is less than or equal to a predefined number of clock cycles  $K_1$  is eliminated. A pulse having a width greater than  $K_1$  clock cycles but less than or equal to  $K_2$  clock cycles is modified to have a width defined by truncating  $K_2$  down to its nearest integer value. A pulse having a width greater than  $K_2$  clock cycles but equal to or less than  $K_3$  clock cycles is modified to have a width defined by truncating  $K_3$  down to its nearest integer value or truncating  $K_2$  up to its nearest integer value. In one embodiment,  $K_1$  may represent a number of clock cycles, e.g., 4. In some embodiments, the truncated pulse width for each range of pulses may be greater than the truncated pulse width of a previous range by one clock cycle.

The above exemplary algorithm may be represented as shown below for one embodiment:

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if  $0 < P \leq K_1$  then  $P = 0$ 
if  $K_1 < P \leq K_2$  then  $P = \text{Integ\_dn}(K_2)$ 
if  $K_2 < P \leq K_3$  then  $P = \text{Integ\_dn}(K_3)$ 
.
.
.
if  $K_i < P \leq K_{i+1}$  then  $P = \text{Integ\_dn}(K_{i+1})$ 

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where function  $\text{Integ\_dn}(\cdot)$  truncates its operand down to its nearest integer value. As described above, the modified pulse widths may also be obtained by either truncating down or

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truncating up either one of the boundaries of each range. In one embodiment,  $\text{Integ\_dn}(K_{i+2})$  is greater than  $\text{Integ\_dn}(K_i)$  by one clock cycle, where  $i$  is an integer greater than or equal to 1.

Assume, for example, that noise shaper **320** operates using a clock frequency of 500 MHz—its period is 2 nsec. In accordance with one exemplary embodiment, noise shaper **320** may be adapted to eliminate pulses that are equal to or less than four clock cycles wide. Pulses that are greater than, for example, 4.0 clock cycles wide but narrower than or equal to, for example, 8.49 clock cycles wide are truncated to have a width of 8 clock cycles; pulses that are greater than, for example, 8.49 clock cycles wide but narrower than or equal to, for example, 9.49 clock cycles wide may be truncated to have a width of 9 clock cycles, etc. The truncation according to this example for four truncation ranges is shown below:

if  $0 < P \leq 4$  then  $P = 0$

if  $4 < P \leq 8.49$  then  $P = 8$

if  $8.49 < P \leq 9.49$  then  $P = 9$

if  $9.49 < P \leq 10.49$  then  $P = 10$

FIG. 9B is a block diagram of one exemplary embodiment of noise shaper **304**. Noise shaper **304** is shown as including, in part, blocks **650**, **652** and **654**. Block **650** is adapted to establish a multitude of ranges each defined by a minimum pulse width characterizing a lower bound of that range and a maximum pulse width characterizing an upper bound of that range. Block **652** is adapted to remove the pulse if the pulse width is narrower than a predefined value. Block **654** is adapted to modify the pulse width to a value characterized by truncating one of the lower or the upper bounds of the range in which the pulse falls.

In one embodiment, block **654** modifies the pulse width to a value characterized by truncating down the upper bound of the range to its nearest integer value. In another embodiment, block **654** modifies the pulse width to a value characterized by truncating up the lower bound of the range to its nearest integer value. In yet another embodiment, block **654** modifies the pulse width to a value characterized by truncating up the upper bound of the range to its nearest integer value. In yet another embodiment, block **654** modifies the pulse width to a value characterized by truncating down the lower bound of the range to its nearest integer value. In yet another embodiment, block **654** modifies the pulse width to an integer number of clock cycles defined by truncating to an integer number either the lower or the upper bound of the range in which the pulse falls.

Referring to FIG. 5, because of the non-zero delay associated with the loop coupling the output terminal of driver **132** to the input terminal of integrator **104**, the pulses generated at the output terminal of driver **132** become extended. The pulse extensions that result from extended integration period by integrator **104** can increase over time leading to degradation of the harmonic distortion of amplifier **100**. To cancel out the loop delay so as to inhibit such pulse extensions, in accordance with one embodiment of the present invention, the reference voltage  $V_{ref}$  applied to comparator **128** is varied, as described further below.

FIG. 12 is a block diagram of a tri-level pulse width modulated amplifier **275** in accordance with another embodiment of the present invention. Positive pulse detector **250** together with phase detector **252**, charge pump **254** and loop filter **256** form a first delay locked-loop **280** adapted to cancel out the

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delay loop that would otherwise cause extended pulses to appear at the output of driver **132**.

Assume a positive pulse, i.e., a pulse making a transition from 0 volt to  $V_{DDP}$  volts, is generated on signal OUT. Such a pulse is detected by positive pulse detector **250**. The width of the pulse generated by pulse width detector **250** is similar to that of the positive pulse generated by output driver **132**.

Phase detector **252** is adapted to compare the phase of the detected pulse P with the phase of the signal M supplied by signal source **102** at node Z. If signal P leads signal M in phase, signal UP generated by phase detector **252** is decreased and signal DN generated by phase detector **252** is increased, thereby causing amplitude of signal S to go down. Conversely, if signal P lags signal M in phase, signal DN generated by phase detector **252** is decreased and signal UP generated by phase detector **252** is increased, thereby causing amplitude of signal S to go up. Loop filter **256** filters out the high frequency components of the signal generated by charge pump **256**. Multiplexer **258** is adapted to enable signal S to be delivered as the reference input voltage  $V_{ref}$  to comparator **128**. Accordingly, the additional voltage that is integrated by integrator **104** as a result of the loop delay—from the output of driver **132** back to the input of integrator **104**—is also tracked by signal  $V_{ref}$ . Consequently, the output voltage of integrator **104** is maintained at nearly 0 volt when a trailing edge is detected on a positive pulse of signal OUT. Delay locked-loop **290** operates in the same manner as delay-locked loop **280** except that it detects the trailing edges of negative pulses of signal OUT and thus maintains the output voltage of integrator **104** at nearly 0 volt when a trailing edge is detected on a negative pulse of signal OUT.

FIG. 13 is an exemplary output voltage of integrator **104**. Absent the loop delay that is subject of the cancellation of the present invention, integrator **104** is adapted to turn off at point  $P_1$ , namely at nearly zero volts. Due to the presence of the loop delay (e.g., 4 nsec), if a positive pulse appears on signal OUT, the integrator's output continues to increase by the shown value of, e.g., 50 mV to point  $P_2$ . To compensate for this increase, delay locked-loop **280** causes voltage  $V_{ref}$  to decrease by 50 mV. Therefore, the effective point at which integrator **104** is turned off remains nearly at nearly zero volts. Similarly, assume a negative pulse appears on signal OUT. Accordingly, the integrator's output continues to decrease by the shown value of, e.g., -50 mV to point  $P_3$ . To compensate for this decrease, delay locked-loop **290** causes voltage  $V_{ref}$  to increase by 50 mV. Therefore, the effective point at which integrator **104** is turned off by driver **132** via the loop remains nearly at zero volts.

FIG. 14 is a block diagram of a tri-level pulse width modulated amplifier **375** in accordance with another embodiment of the present invention. Positive edge detector **380** together with sample-and-hold circuit **390**, comparator **384**, charge pump **382** and capacitor **392** form a first cancellation loop adapted to cancel out the loop delays that would otherwise cause pulse extensions to appear at the output of driver **132**. Similarly, negative edge detector **480** together with sample-and-hold circuit **490**, comparator **484**, charge pump **482** and capacitor **492** form a second cancellation loop adapted to cancel out the loop delays that would otherwise cause pulse extensions to appear at the output of driver **132**. Positive edge detector **380** is adapted to generate a relatively short positive pulse in response to detecting a trailing edge of a positive pulse on signal OUT, i.e., in response to detecting a signal OUT edge making a transition from  $V_{DDP}$  to 0. Similarly, negative edge detector **480** is adapted to generate a relatively short positive pulse in response to detecting a trailing edge of

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a negative pulse on signal OUT, i.e., in response to detecting a signal OUT edge making a transition from VDDN to 0.

When in response to a trailing edge of a positive pulse of signal OUT, positive edge detector 380 generates a positive pulse, the voltage present at the output of integrator 104 is sampled by transistor 386 and held by capacitor 388 of sample-and-hold circuit 390. The voltage across capacitor 388 is compared by comparator 384 that receives signal Vss (e.g., ground potential) as its reference voltage. If the output voltage of integrator 104 is greater than Vss, the output of comparator 384 goes high, thereby causing charge pump 382 to discharge capacitor 392. Conversely, if the output voltage of integrator 104 is smaller than Vss, the output of comparator 384 goes low, thereby causing charge pump 382 to charge capacitor 390. Multiplexer 395 is adapted to cause the voltage at the output of charge pump 382 to be delivered as the reference voltage to comparator 128 when a positive pulse is detected on signal OUT. The cancellation loop as described above causes the output voltage of integrator 104 to be nearly at zero volts when a trailing edge is detected on a positive signal OUT.

When in response to a trailing edge of a negative pulse of signal OUT, negative edge detector 480 generates a positive pulse, the voltage present at the output of integrator 104 is sampled by transistor 486 and held by capacitor 488 of sample-and-hold circuit 490. The voltage across capacitor 488 is compared by comparator 484 that receives signal Vss as its reference voltage. If the output voltage of integrator 104 is greater than Vss, the output of comparator 484 goes high, thereby causing charge pump 482 to discharge capacitor 490. Conversely, if the output voltage of integrator 104 is smaller than Vss, the output of comparator 484 goes low, thereby causing charge pump 482 to charge capacitor 490. Multiplexer 395 is adapted to cause the voltage at the output of charge pump 492 to be delivered as the reference voltage to comparator 128 when a negative pulse is detected on signal OUT. The cancellation loop thus causes the output voltage of integrator 104 to be nearly at zero volts when a trailing edge is detected on a negative signal OUT.

FIG. 15A is an exemplary timing diagram of a number of signals associated with amplifier 100, shown in FIG. 5. In accordance with this timing diagram, a positive pulse from 0 to VDDP is generated by output driver 132, i.e., signal OUT. The process of generating the positive pulse begins at time 200. At time 200, pulse width modulator 306 causes signal POS signal to go high, thereby causing switch 170 to close. Signal NEG remains low to keep switch 168 open. The width (duration) of pulse signal POS is determined based on data  $D_3$  received from noise shaper 304, as described above. After the elapse of time period  $T_0$ , pulse width modulator 306 causes signal POS\_DELAY to make a low-to-high transition at time 202. During the period  $T_0$ , the output voltage of driver 132 is at the ground potential, therefore, integrator 104 integrates only the signal it receives from signal source 102 and its output increases during this time. Because signal POS is at a high level during this time, the voltage supplied to integrator 104 via node Z is supplied by voltage divider 180 and op-amp 166.

When signal POS\_DELAY makes a low-to-high transition at time 202, signals PG, NG and OG go low. This causes transistor 220 to turn on and transistors 222 and 224 to be off, in turn causing output signal OUT carried by node A to be pulled to VDDP. The high voltage at node A is applied to integrator 104 via feedback resistor 122. From time 202, when signals POS\_DELAY makes a low-to-high transition, until time 204, when signal POS makes a high-to-low transition, integrator 104 integrates the difference between the

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signals supplied by signal source 102 at node Z, and the signal received from feedback resistor 122. If this difference is zero, the output of integrator 104 remains generally flat as shown at 210. If this difference is positive, the output of integrator 104 increases as shown at 212. If this difference is negative, the output of integrator 104 decreases as shown at 214.

Pulse width modulator 306 returns signals POS and POS\_DELAY to zero at times 204 and 206, respectively. In some embodiments, the delay  $T_1$ —between the times 206 and 204—is twice the delay  $T_0$ —between the times 202 and 200. In some embodiments, delays  $T_0$  and  $T_1$  are fractions of the width of signal POS. Time delays  $T_0$  and  $T_1$  provide integrator 104 with sufficient time to integrate and discharge, respectively, without causing comparator 128 to generate a multitude of signal transitions, or to go into saturation. Saturation refers to a condition according to which the integrator's output tries either to reach or exceed the maximum possible output voltage or to reach or go below a minimum possible output voltage.

At time 204 the output of integrator 104 begins to fall at decay rate 216. The duration of the decay depends on the level of the integrator 104 output voltage signal at time 204. At time 218 the output signal from integrator 104 is assumed to cross the constant reference signal of comparator 128. The output of comparator 128 thus switches, in turn causing signal OUT generated by driver 132 to be lowered from VDDP to the ground potential. Occurrence of time 218 is dependent on the level of output signal of integrator 104 at time 204 and decay rate 216. Integrator 104 therefore adjusts the duration of signal OUT to accommodate for fluctuations in the supply voltages VDDP, VDDN, the voltage drops across transistor switches 220 and 224 of driver 132, and overshoots and undershoots of signal OUT. Integrator 104 therefore adjusts the duration of signal OUT such that the area covered by signal OUT is approximately constant for a given  $D_3$  received from noise shaper 176.

FIG. 15B is an exemplary timing diagram of a number of signals associated with amplifier 100, shown in FIG. 5. In accordance with this timing diagram, a negative pulse from 0 to VDDN is generated by output driver 132. The process of generating the negative pulse begins at time 300. At time 300, pulse-width modulator 306 causes signal NEG to go high, thereby causing switch 168 to close. Signal POS remains low to keep switch 170 open. The width (duration) of pulse signal NEG is determined based on data  $D_3$  received from noise shaper 304, as described above. After an elapse of time period  $T_2$ , pulse-width modulator 306 causes signal NEG\_DELAY to make a low-to-high transition at time 302. During the period  $T_2$ , the output voltage of driver 132 is at the ground potential, therefore, integrator 104 integrates only the signal it receives from signal source 102, therefore its output voltage decreased during this time. Because signal NEG is at a high level during this time, the voltage supplied to integrator 104 via node Z is supplied by voltage divider 175 and op-amp 164.

When signal NEG\_DELAY makes a low-to-high transition at time 302, signal OG goes low and signal NG goes high. This causes transistor 222 to turn off and transistor 224 to turn on, in turn causing output signal OUT carried by node A to be pulled low to negative supply voltage VDDN. The voltage VDDN at node A is applied to integrator 104 via feedback resistor 122. From time 302, when signals NEG\_DELAY makes a low-to-high transition, until time 304, when signal NEG makes a high-to-low transition, integrator 104 integrates the difference between the signals supplied by signal source 102 at node Z, and the signal received from feedback resistor 122. If this difference is zero, the output of integrator 104 remains generally flat as shown at 310. If this difference

is positive, the output of integrator **104** decreases as shown at **312**. If this difference is negative, the output of integrator **104** decreases as shown at **314**.

Pulse width modulator **306** returns signals NEG and NEG\_DELAY to zero at times **304** and **306**, respectively. In some embodiment, the delay  $T_3$ —between the times **306** and **304**—is twice the delay  $T_2$ —between the times **302** and **300**. In some embodiments, delays  $T_2$  and  $T_3$  are fractions of the width of signal NEG. Time delays  $T_2$  and  $T_3$  provide integrator **104** with sufficient time to integrate and discharge, respectively, without causing comparator **128** to generate a multitude of signal transitions.

At time **304** the output of integrator **104** begins to rise at a decay rate **316**. The duration of the rise depends on the level of the integrator **104** output voltage signal at time **304**. At time **318** the output signal from integrator **104** is assumed to cross the constant reference signal of comparator **128**. The output of comparator **128** thus switches, in turn causing signal OUT generated by driver **132** to be raised from VDDN to the ground potential. Occurrence of time **318** is dependent on the level of output signal of integrator **104** at time **304** and decay rate **316**. Integrator **104** therefore adjusts the duration of signal OUT to accommodate for fluctuations in the supply voltages VDDP, VDDN, and the voltage drops across transistor switches **220** and **224** of driver **132**. Integrator **104** therefore adjusts the duration of signal OUT such that the area covered by the duration of signal OUT is approximately constant for a given signal  $D_3$  received from noise shaper **176**.

FIGS. **16A** and **16B** show exemplary timing relationship between signal NEG\_DELAY, the output signal of comparator **128**, CMP\_OUT, and signal PG applied to the gate terminal of transistor **220** disposed in driver **132**. During a predefined period  $T_8$  while signal NEG\_DELAY remains asserted, in response to a low-to-high transition **600** of signal CMP\_OUT, signal PG is asserted, e.g., makes a high-to-low transition **602**. Signal PG remains asserted until one of two conditions is satisfied. First, referring to FIG. **16A**, signal PG is deasserted—shown using transition **606**—if signal CMP\_OUT makes a high-to-low transition **604** during the period  $T_8$ . When signal CMP\_OUT makes such a high-to-low transition, a pulse generated within switch pulse logic **130** deasserts signal PG. Second, referring to FIG. **16B**, if a high-to-low transition on signal CMP\_OUT does not occur during period  $T_8$ , signal PG remains asserted until signal NEG\_DELAY makes a high-to-low transition **608**, in response to which signal PG is deasserted—shown using transition **606**.

FIGS. **17A** and **17B** show exemplary timing relationship between signal NEG\_DELAY, the output signal of comparator **128**, CMP\_OUT, and signal NG applied to the gate terminal of transistor **224** disposed in driver **132**. During a predefined period  $T_9$  while signal NEG\_DELAY remains asserted, in response to a high-to-low transition **700** of signal CMP\_OUT, signal NG is asserted, e.g., makes a low-to-high transition **702**. Signal NG remains asserted until one of two conditions is satisfied. First, referring to FIG. **16B**, signal NG is deasserted—shown using transition **706**—if signal CMP\_OUT makes a high-to-low transition **704** during period  $T_9$ . When signal CMP\_OUT makes such a low-to-high transition, a pulse generated within switch pulse logic **130** deasserts signal NG. Second, referring to FIG. **17B**, if a low-to-high transition on signal CMP\_OUT does not occur during period  $T_9$ , signal NG remains asserted until signal NEG\_DELAY makes a high-to-low transition **708**, in response to which signal NG is deasserted—shown using transition **706**.

FIG. **18** is a block diagram of switch pulse logic **130**. Switch pulse logic **130** is shown as including a pair of one-shot logic blocks **822**, **824** and inverter **826**. Signal CMP\_OUT generated by comparator **128** is applied to the input terminal CMP of one-shot block **824**. Inverter **826** applies the inverse of signal CMP\_OUT to the input terminal

CMP of one-shot block **822**. Signals POS\_DELAY and NEG\_DELAY are applied to the input terminal EN of one-shot logic blocks **822** and **824** respectively. One-shot logic blocks **822** and **824** are adapted to suppress spikes that may appear on signal CMP\_OUT. Furthermore, the pulse width of the signal CMP\_OUT is increased by a minimum predefined value if it is detected by the one-shot logic blocks **822** and **824** to be shorter than a pre-established value.

FIG. **19** shows the various components of each of one-shot logic blocks **822** and **824**, in accordance with one embodiment of the present invention. Each of these one-shot logic blocks is shown as including an AND gate **830**, a NAND gate **832**, set-reset latches **834**, **836**, and pulse generators **838** and **840**. Signal Q of each of latches **834** and **836** is set to 1 if the input signal SB of that latch is set to zero. Similarly, signal Q of each of latches **834** and **836** is set to zero if the input signal RB of that latch is set to zero. Each of pulse generators **838** and **840** generates a one-shot pulse of a predefined duration when it receives a transition on its input signal IN.

Referring now to FIGS. **20A-20G**, various exemplary implementations of the present invention are shown. Referring to FIG. **20A**, the present invention may be embodied in a hard disk drive **900**. The present invention may implement either or both signal processing and/or control circuits, which are generally identified in FIG. **20A** at **902**. In some implementations, signal processing and/or control circuit **902** and/or other circuits (not shown) in HDD **900** may process data, perform coding and/or encryption, perform calculations, and/or format data that is output to and/or received from a magnetic storage medium **906**.

HDD **900** may communicate with a host device (not shown) such as a computer, mobile computing devices such as personal digital assistants, cellular phones, media or MP3 players and the like, and/or other devices via one or more wired or wireless communication links **908**. HDD **900** may be connected to memory **909**, such as random access memory (RAM), a nonvolatile memory such as flash memory, read only memory (ROM) and/or other suitable electronic data storage.

Referring now to FIG. **20B**, the present invention may be embodied in a digital versatile disc (DVD) drive **910**. The present invention may implement either or both signal processing and/or control circuits, which are generally identified in FIG. **20B** at **912**, and/or mass data storage **918** of DVD drive **910**. Signal processing and/or control circuit **912** and/or other circuits (not shown) in DVD **910** may process data, perform coding and/or encryption, perform calculations, and/or format data that is read from and/or data written to an optical storage medium **916**. In some implementations, signal processing and/or control circuit **912** and/or other circuits (not shown) in DVD **910** can also perform other functions such as encoding and/or decoding and/or any other signal processing functions associated with a DVD drive.

DVD drive **910** may communicate with an output device (not shown) such as a computer, television or other device via one or more wired or wireless communication links **917**. DVD **910** may communicate with mass data storage **918** that stores data in a nonvolatile manner. Mass data storage **918** may include a hard disk drive (HDD) such as that shown in FIG. **20A**. The HDD may be a mini HDD that includes one or more platters having a diameter that is smaller than approximately 1.8". DVD **910** may be connected to memory **919**, such as RAM, ROM, nonvolatile memory such as flash memory, and/or other suitable electronic data storage.

Referring now to FIG. **20C**, the present invention may be embodied in a high definition television (HDTV) **920**. The present invention may implement either or both signal processing and/or control circuits, which are generally identified in FIG. **20C** at **922**, a WLAN interface **929** and/or mass data storage of the HDTV **920**. HDTV **920** receives HDTV input

signals in either a wired or wireless format and generates HDTV output signals for a display 926. In some implementations, signal processing circuit and/or control circuit 922 and/or other circuits (not shown) of HDTV 920 may process data, perform coding and/or encryption, perform calculations, format data and/or perform any other type of HDTV processing that may be required.

HDTV 920 may communicate with mass data storage 927 that stores data in a nonvolatile manner such as optical and/or magnetic storage devices. At least one HDD may have the configuration shown in FIG. 20A and/or at least one DVD may have the configuration shown in FIG. 20B. The HDD may be a mini HDD that includes one or more platters having a diameter that is smaller than approximately 1.8". HDTV 920 may be connected to memory 928 such as RAM, ROM, nonvolatile memory such as flash memory and/or other suitable electronic data storage. HDTV 920 also may support connections with a WLAN via a WLAN network interface 929.

Referring now to FIG. 20D, the present invention implements a control system of a vehicle 930, a WLAN interface 948 and/or mass data storage of the vehicle control system. In some implementations, the present invention implements a powertrain control system 932 that receives inputs from one or more sensors such as temperature sensors, pressure sensors, rotational sensors, airflow sensors and/or any other suitable sensors and/or that generates one or more output control signals such as engine operating parameters, transmission operating parameters, and/or other control signals.

The present invention may also be embodied in other control systems 940 of vehicle 930. Control system 940 may likewise receive signals from input sensors 942 and/or output control signals to one or more output devices 944. In some implementations, control system 940 may be part of an anti-lock braking system (ABS), a navigation system, a telematics system, a vehicle telematics system, a lane departure system, an adaptive cruise control system, a vehicle entertainment system such as a stereo, DVD, compact disc and the like. Still other implementations are contemplated.

Powertrain control system 932 may communicate with mass data storage 946 that stores data in a nonvolatile manner. Mass data storage 946 may include optical and/or magnetic storage devices for example hard disk drives HDD and/or DVDs. At least one HDD may have the configuration shown in FIG. 20A and/or at least one DVD may have the configuration shown in FIG. 20B. The HDD may be a mini HDD that includes one or more platters having a diameter that is smaller than approximately 1.8". Powertrain control system 932 may be connected to memory 947 such as RAM, ROM, nonvolatile memory such as flash memory and/or other suitable electronic data storage. Powertrain control system 932 also may support connections with a WLAN via a WLAN network interface 948. The control system 940 may also include mass data storage, memory and/or a WLAN interface (all not shown).

Referring now to FIG. 20E, the present invention may be embodied in a cellular phone 950 that may include a cellular antenna 951. The present invention may implement either or both signal processing and/or control circuits, which are generally identified in FIG. 20E at 952, a WLAN interface 968 and/or mass data storage 964 of the cellular phone 950. In some implementations, cellular phone 950 includes a microphone 956, an audio output 958 such as a speaker and/or audio output jack, a display 960 and/or an user input 962 such as a keypad, pointing device, voice actuation and/or other input device. Signal processing and/or control circuits 952 and/or other circuits (not shown) in cellular phone 950 may process data, perform coding and/or encryption, perform calculations, format data and/or perform other cellular phone functions.

Cellular phone 950 may communicate with mass data storage 964 that stores data in a nonvolatile manner such as optical and/or magnetic storage devices for example hard disk drives HDD and/or DVDs. At least one HDD may have the configuration shown in FIG. 20A and/or at least one DVD may have the configuration shown in FIG. 20B. The HDD may be a mini HDD that includes one or more platters having a diameter that is smaller than approximately 1.8". Cellular phone 950 may be connected to memory 966 such as RAM, ROM, nonvolatile memory such as flash memory and/or other suitable electronic data storage. Cellular phone 950 also may support connections with a WLAN via a WLAN network interface 968.

Referring now to FIG. 20F, the present invention may be embodied in a set top box 980. The present invention may implement either or both signal processing and/or control circuits, which are generally identified in FIG. 20F at 984, a WLAN interface 996 and/or mass data storage 990 of the set top box 980. Set top box 980 receives signals from a source such as a broadband source and outputs standard and/or high definition audio/video signals suitable for a display 988 such as a television and/or monitor and/or other video and/or audio output devices. Signal processing and/or control circuits 984 and/or other circuits (not shown) of the set top box 980 may process data, perform coding and/or encryption, perform calculations, format data and/or perform any other set top box function.

Set top box 480 may communicate with mass data storage 990 that stores data in a nonvolatile manner. Mass data storage 990 may include optical and/or magnetic storage devices for example hard disk drives HDD and/or DVDs. At least one HDD may have the configuration shown in FIG. 20A and/or at least one DVD may have the configuration shown in FIG. 20B. The HDD may be a mini HDD that includes one or more platters having a diameter that is smaller than approximately 1.8". Set top box 980 may be connected to memory 994 such as RAM, ROM, nonvolatile memory such as flash memory and/or other suitable electronic data storage. Set top box 980 also may support connections with a WLAN via a WLAN network interface 996.

Referring now to FIG. 20G, the present invention may be embodied in a media player 972. The present invention may implement either or both signal processing and/or control circuits, which are generally identified in FIG. 20G at 971, a WLAN interface 974 and/or mass data storage of the media player 972. In some implementations, media player 972 includes a display 976 and/or a user input 977 such as a keypad, touchpad and the like. In some implementations, media player 972 may employ a graphical user interface (GUI) that typically employs menus, drop down menus, icons and/or a point-and-click interface via display 976 and/or user input 977. Media player 972 further includes an audio output 975 such as a speaker and/or audio output jack. Signal processing and/or control circuits 971 and/or other circuits (not shown) of media player 972 may process data, perform coding and/or encryption, perform calculations, format data and/or perform any other media player function.

Media player 972 may communicate with mass data storage 970 that stores data such as compressed audio and/or video content in a nonvolatile manner. In some implementations, the compressed audio files include files that are compliant with MP3 format or other suitable compressed audio and/or video formats. The mass data storage may include optical and/or magnetic storage devices for example hard disk drives HDD and/or DVDs. At least one HDD may have the configuration shown in FIG. 20A and/or at least one DVD may have the configuration shown in FIG. 20B. The HDD may be a mini HDD that includes one or more platters having a diameter that is smaller than approximately 1.8". Media player 972 may be connected to memory 973 such as RAM,

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ROM, nonvolatile memory such as flash memory and/or other suitable electronic data storage. Media player 972 also may support connections with a WLAN via a WLAN network interface 974.

Referring to FIG. 20H, the present invention may be embodied in a Voice over Internet Protocol (VoIP) phone 983 that may include an antenna 939. The present invention may implement either or both signal processing and/or control circuits, which are generally identified in FIG. 20H at 982, a wireless interface and/or mass data storage of the VoIP phone 983. In some implementations, VoIP phone 983 includes, in part, a microphone 987, an audio output 989 such as a speaker and/or audio output jack, a display monitor 991, a user input 992 such as a keypad, pointing device, voice actuation and/or other input devices, and a Wireless Fidelity (Wi-Fi) communication module 986. Signal processing and/or control circuits 982 and/or other circuits (not shown) in VoIP phone 983 may process data, perform coding and/or encryption, perform calculations, format data and/or perform other VoIP phone functions.

VoIP phone 983 may communicate with mass data storage 502 that stores data in a nonvolatile manner such as optical and/or magnetic storage devices, for example hard disk drives HDD and/or DVDs. At least one HDD may have the configuration shown in FIG. 20A and/or at least one DVD may have the configuration shown in FIG. 20B. The HDD may be a mini HDD that includes one or more platters having a diameter that is smaller than approximately 1.8". VoIP phone 983 may be connected to memory 985, which may be a RAM, ROM, nonvolatile memory such as flash memory and/or other suitable electronic data storage. VoIP phone 983 is configured to establish communications link with a VoIP network (not shown) via Wi-Fi communication module 986. Still other implementations in addition to those described above are contemplated. Although not explicitly shown, the present invention may also be disposed in various other portable electronic devices, such as MP3 players, personal digital assistants (PDA), notebook computers, etc.

The above embodiments of the present invention are illustrative and not limitative. Various alternatives and equivalents are possible. The invention is not limited by the type of integrated circuit in which the present disclosure may be disposed. Nor is the invention limited to any specific type of process technology, e.g., CMOS, Bipolar, or BICMOS that may be used to manufacture the present disclosure. Other additions, subtractions or modifications are obvious in view of the present disclosure and are intended to fall within the scope of the appended claims.

What is claimed is:

1. A method of truncating a pulse having a width, the method comprising:

establishing a plurality of ranges each range defined by a minimum pulse width characterizing a lower bound of that range and a maximum pulse width characterizing an upper bound of that range;

removing the pulse if the pulse width is narrower than a predefined value; and

modifying the pulse width to a value characterized by truncating one of the lower or the upper bounds of the range in which the pulse falls.

2. The method of claim 1 wherein the step of truncating comprises:

modifying the pulse width to a value characterized by truncating down the upper bound of the range to its nearest integer value.

3. The method of claim 1 wherein the step of truncating comprises:

modifying the pulse width to a value characterized by truncating up the lower bound of the range to its nearest integer value.

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4. The method of claim 1 wherein the step of truncating comprises:

modifying the pulse width to a value characterized by truncating up the upper bound of the range to its nearest integer value.

5. The method of claim 1 wherein the step of truncating comprises:

modifying the pulse width to a value characterized by truncating down the lower bound of the range to its nearest integer value.

6. The method of claim 1 wherein a difference between the predefined value and a truncated upper bound of a succeeding range is greater than 1 clock cycle.

7. The method of claim 1 further comprising:

modifying the pulse width to an integer number of clock cycles defined by truncating to the integer number one of the lower or the upper bounds of the range in which the pulse falls.

8. A noise shaper adapted to truncate a pulse having a pulse width, the noise shaper comprising:

a block adapted to establish a plurality of ranges each range defined by a minimum pulse width characterizing a lower bound of that range and a maximum pulse width characterizing an upper bound of that range;

a block adapted to remove the pulse if the pulse width is narrower than a predefined value; and

a block adapted modify the pulse width to a value characterized by truncating one of the lower or the upper bounds of the range in which the pulse falls.

9. The noise shaper of claim 8 wherein said modifying block is further adapted to modify the pulse width to a value characterized by truncating down the upper bound of the range to its nearest integer value.

10. The noise shaper of claim 8 wherein said modifying block is further adapted to modify the pulse width to a value characterized by truncating up the lower bound of the range to its nearest integer value.

11. The noise shaper of claim 8 wherein said modifying block is further adapted to modify the pulse width to a value characterized by truncating up the upper bound of the range to its nearest integer value.

12. The noise shaper of claim 8 wherein said modifying block is further adapted to modify the pulse width to a value characterized by truncating down the lower bound of the range to its nearest integer value.

13. The noise shaper of claim 8 wherein a difference between the predefined value and a truncated upper bound of a succeeding range is greater than 1 clock cycle.

14. The noise shaper of claim 8 wherein said modifying block is further adapted to modify the pulse width to an integer number of clock cycles defined by truncating to the integer number one of the lower bound or the upper bound of the range in which the pulse falls.

15. The noise shaper of claim 8 wherein said noise shaper is disposed in an amplifier.

16. The noise shaper of claim 15 wherein said amplifier is adapted to amplify audio signals.

17. The noise shaper of claim 15 wherein said amplifier is adapted to amplify video signals.

18. A circuit comprising:

an integrator adapted to integrate a first signal in accordance with a feedback signal;

a comparator responsive to an output of the integrator and a reference voltage;

a first pulse detector adapted to detect a pulse;

a first phase detector responsive to an output of the first pulse detector and the first signal; and

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a first charge pump responsive to the first phase detector and adapted to generate the reference voltage selectively applied to the comparator.

19. The circuit of claim 18 wherein said pulse is a positive pulse, the circuit further comprising:

a second pulse detector adapted to detect a negative pulse;

a second phase detector responsive to an output of the second pulse detector and the first signal;

a second charge pump responsive to the second phase detector and adapted to generate a second reference voltage; and

a multiplexer adapted to select between the first and second reference voltages to supply a selected reference voltage and apply the selected reference voltage to the comparator.

20. The circuit of claim 19 further comprising:  
a driver adapted to generate the positive pulse and said negative pulse at its output terminal.

21. The circuit of claim 20 further comprising:  
a logic block adapted to control the driver in response to first and second pulse-width modulated signals.

22. The circuit of claim 20 wherein said circuit is disposed in an amplifier.

23. The circuit of claim 22 wherein said amplifier is adapted to amplify audio signals.

24. The circuit of claim 22 wherein said amplifier is adapted to amplify video signals.

25. A method of minimizing effect of a loop delay on an output signal, the method comprising:

integrating a first signal in accordance with a feedback signal to generate an integrated signal;

comparing the integrated signal to a first reference signal;

detecting a pulse;

detecting a difference between a phase of the pulse and a phase of the first signal; and

varying the first reference signal in response to the detected phase difference.

26. The method of claim 25 wherein said pulse is a positive pulse, the method further comprising:

detecting a negative pulse;

detecting a difference between a phase of the negative pulse and a phase of the first signal; and

varying the first reference signal in response to the detected difference between the phase of the negative pulse and the phase of the first signal.

27. The method of claim 26 wherein said positive pulse varies between a ground potential and a positive supply voltage and wherein said negative pulse varies between the ground potential and a negative supply voltage.

28. The method of claim 27 further comprising:  
generating said positive and negative pulses in response to first and second pulse-width modulated signals.

29. A circuit comprising:  
an integrator adapted to integrate a first signal in accordance with a feedback signal;

a comparator responsive to an output of the integrator and a first reference voltage;

a first edge detector adapted to detect a trailing edge of the feedback signal;

a first sample-and-hold circuit responsive to an output of the integrator and the detected trailing edge;

a second comparator responsive to an output of the first sample-and-hold circuit and to a supply potential; and

a first charge pump responsive to an output of the second comparator and adapted to vary the first reference voltage selectively applied to the first comparator.

30. The circuit of claim 29 wherein said first edge detector is a positive edge detector, the circuit further comprising:

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a negative edge detector adapted to detect a trailing edge of a negative pulse of the feedback signal;

a second sample-and-hold circuit responsive to the output of the integrator;

a third comparator responsive to an output of the second sample-and-hold circuit and to the supply potential;

a second charge pump responsive an output of the third comparator and adapted to vary a second reference voltage; and

a multiplexer adapted to select between the first and second reference voltages to supply a selected reference voltage and apply the selected reference voltage to the first comparator.

31. The circuit of claim 30 wherein said supply potential is a ground potential.

32. The circuit of claim 30 further comprising:  
a logic block adapted to generate first, second and third control signals in response to an output of the first comparator and further in response to first and second pulse width modulated signals.

33. The circuit of claim 32 further comprising:  
a driver adapted to generate the feedback signal in response to said first, second and third control signals.

34. The circuit of claim 33 wherein said circuit is disposed in an amplifier.

35. The circuit of claim 34 wherein said amplifier is adapted to amplify audio signals.

36. The circuit of claim 34 wherein said amplifier is adapted to amplify video signals.

37. A method of minimizing effect of a loop delay on an output signal:

integrating a first signal in accordance with a feedback signal;

comparing the integrated signal to a first reference signal;

detecting a trailing edge of the feedback signal;

sampling the integrated signal in response to the detected trailing edge;

holding the sampled signal;

comparing the held signal to a supply potential to generate a first compare signal; and

varying the first reference signal in response to the first compare signal.

38. The method of claim 37 wherein said trailing edge is a trailing edge of a positive pulse, the method further comprising:

detecting a trailing edge of a negative pulse of the feedback signal;

sampling the integrated signal in response to the detected trailing edge of the negative pulse;

holding the signal sampled in response to the detected trailing edge of the negative pulse;

comparing the signal held in response to the detected trailing edge of the negative pulse to the supply potential to generate a compare signal; and

varying the first reference signal in response to the second compare signal.

39. The method of claim 38 wherein said supply potential is a ground potential.

40. The method of claim 38 wherein said positive pulse varies between a ground potential and a positive supply voltage and wherein said negative pulse varies between the ground potential and a negative supply voltage.

41. The method of claim 40 further comprising:  
generating said feedback in response to first and second pulse-width modulated signals.